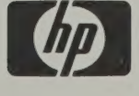
Measurements of High Frequency Networks

SANTA ROSA
NETWORK MEASUREMENTS DIVISION



HEWLETT PACKARD

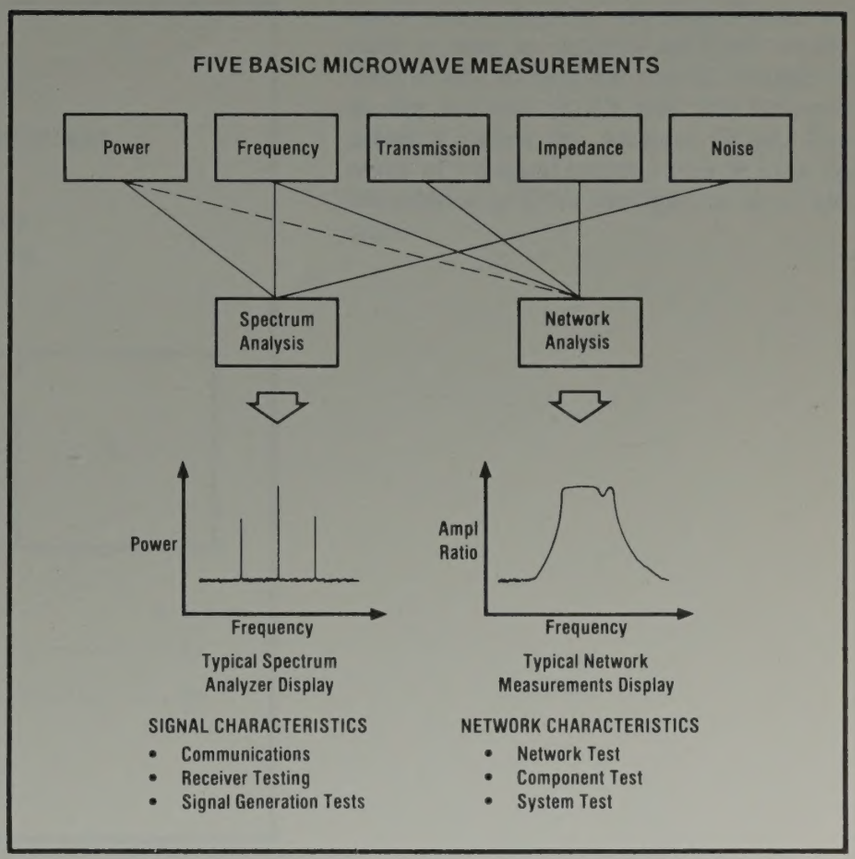
Of the five basic microwave measurements, Network Analysis is primarily concerned with three: frequency, transmission and impedance.

Transmission: how efficiently energy is transferred through the device

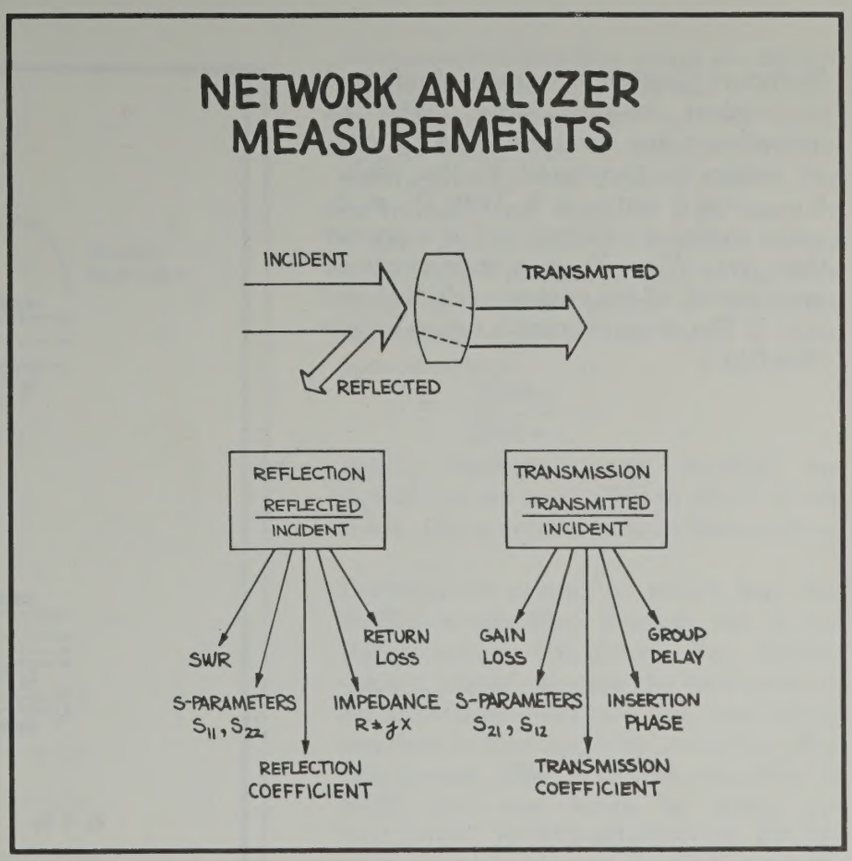
Impedance: how effectively energy is transferred into the device

Frequency: Usually parameters are measured as a function of frequency but only to the accuracy of the source calibration.

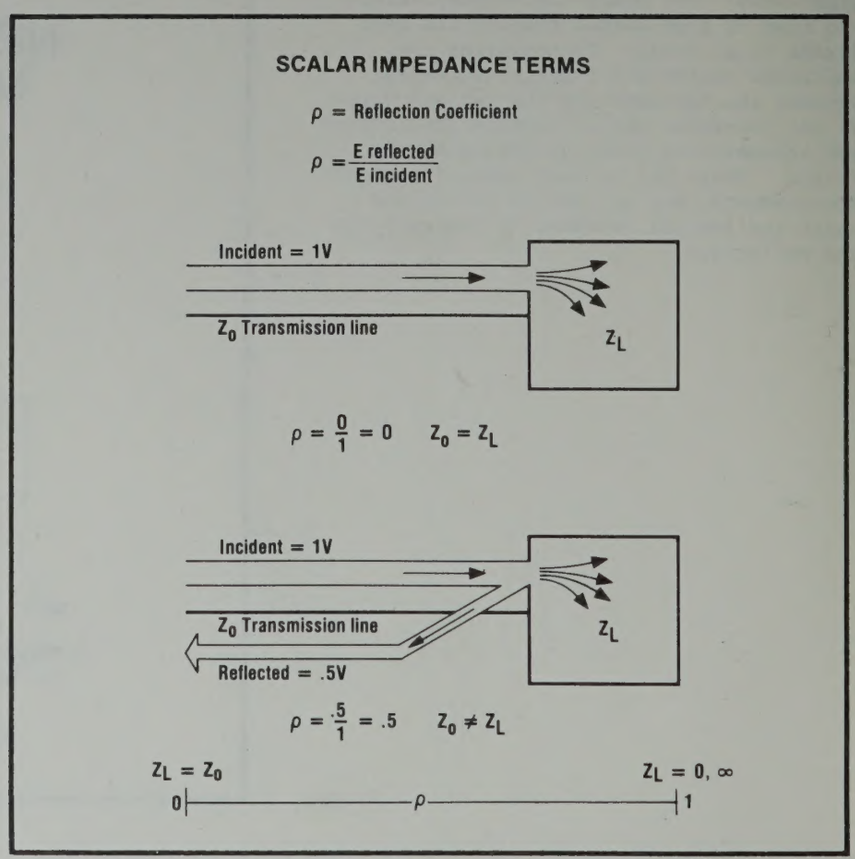
Spectrum Analysis is primarily concerned with characterizing a signal; i.e., its spurious and harmonic components, modulation, noise, etc. It indicates discrete frequencies where microwave energy exists.

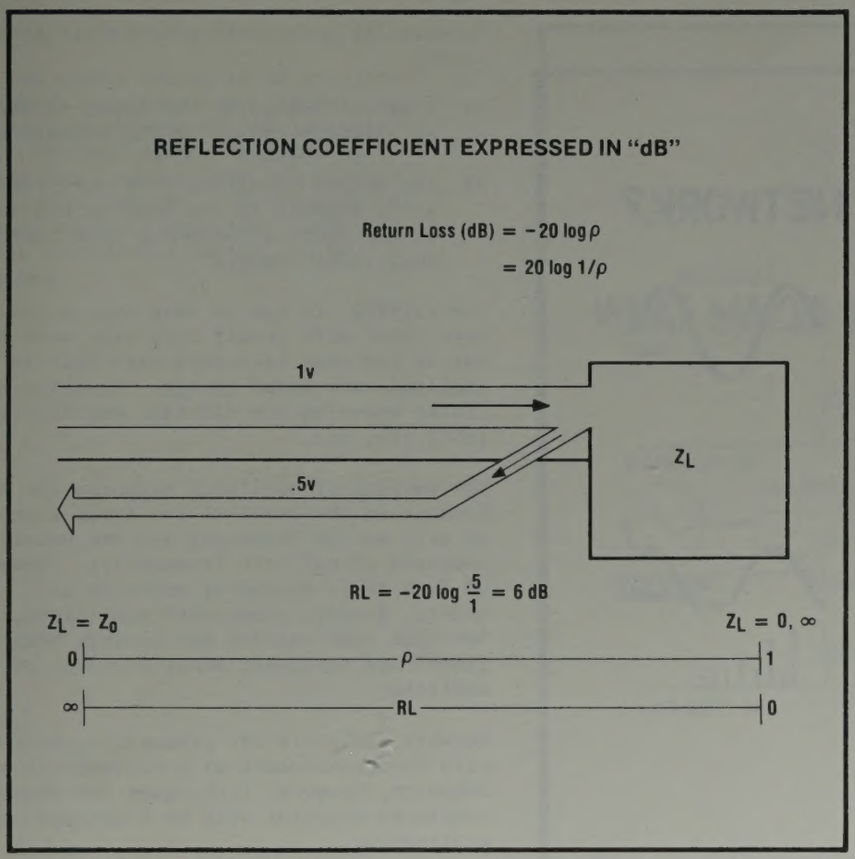


RF or microwave energy can be likened to a light wave. The energy is either reflected from or transmitted through the test device (e.g. lens). By measuring the amplitude ratios and phase differences between the incident and the two new waves we can determine the reflection (impedance) and transmission characteristics of the device. There may be many names for these measurements, but all can be classified under the general headings of transmission and reflection.

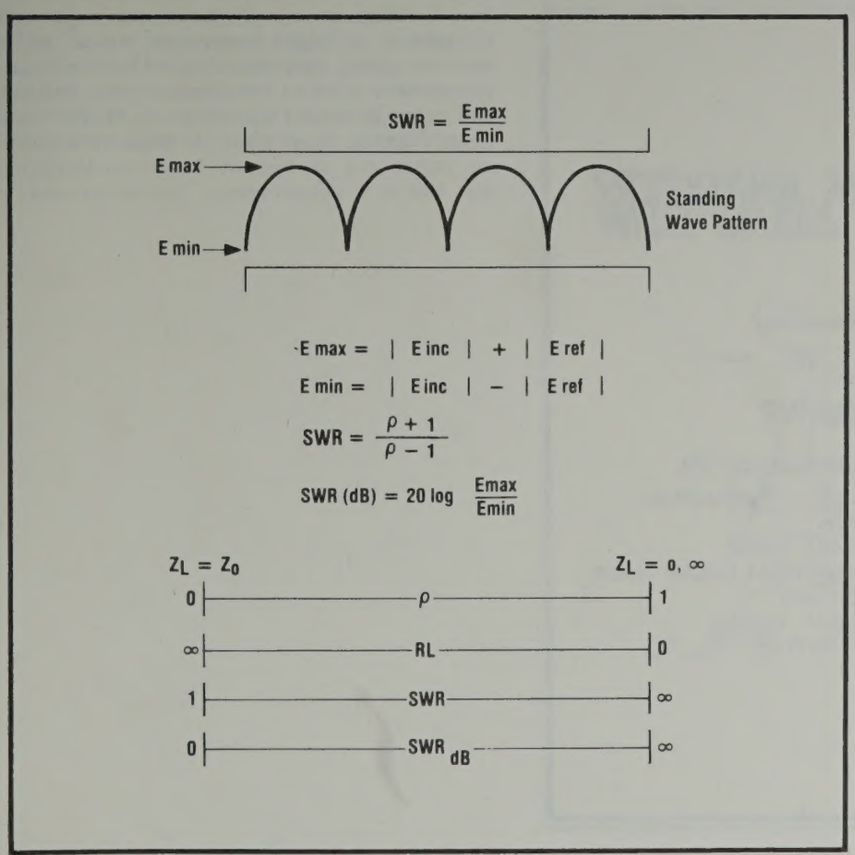


Reflection Coefficient = ratio of reflected to incident traveling wave. With Z_O transmission line terminated in a Z_O load, all energy is transferred to the load: $E_{reflected} = 0$ and $\rho = 0$. With $Z_L \neq Z_O$ some energy is reflected and ρ is greater than zero. When Z_L = a short circuit or open circuit, all the energy is reflected and $\rho = 1$. The range of possible values for ρ is then 0 to 1.





Since many displays are logarithmic, we need a term to express reflection coefficient in dB. Return loss can be thought of as the number of dB that the reflected signal is below the incident signal. The range of values of return loss is ∞ for a Z_O impedance to 0 for an open or short circuit.

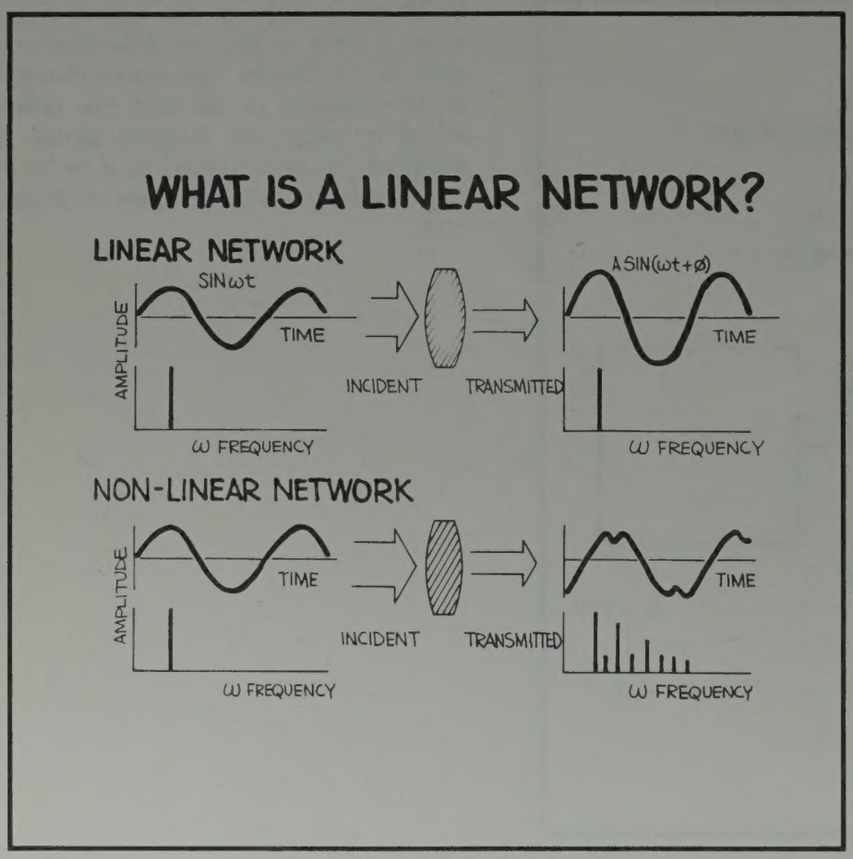


Whenever two traveling waves are present on a transmission line, they add vectorially to produce a stationary standing wave. E_{max} and E_{min} occur at different places on the transmission line where E_r and E_i add (in phase) and subtract (180° out of phase). These values are obtained using a slotted line. The ratio of E_{max} to E_{min} then gives us another means of evaluating relative impedance. This value is easily converted to ρ:

$$\rho = \frac{SWR-1}{SWR+1}$$

Again, because many displays are logarithmic we have SWR in dB = 20 log SWR. This is *not* the same as Return Loss.

Conversions among ρ, return loss and SWR are simplified through use of the Reflectometer Calculator. The known value is placed opposite the blue arrow in the appropriate window. The other values can then be read opposite one of the other blue arrows. SWR_{dB} can be converted to SWR and vice versa by using the VOLTAGE RATIO-dB-POWER RATIO scale at the bottom of the Reflectometer Calculator.



Technically a network's behavior is linear where:

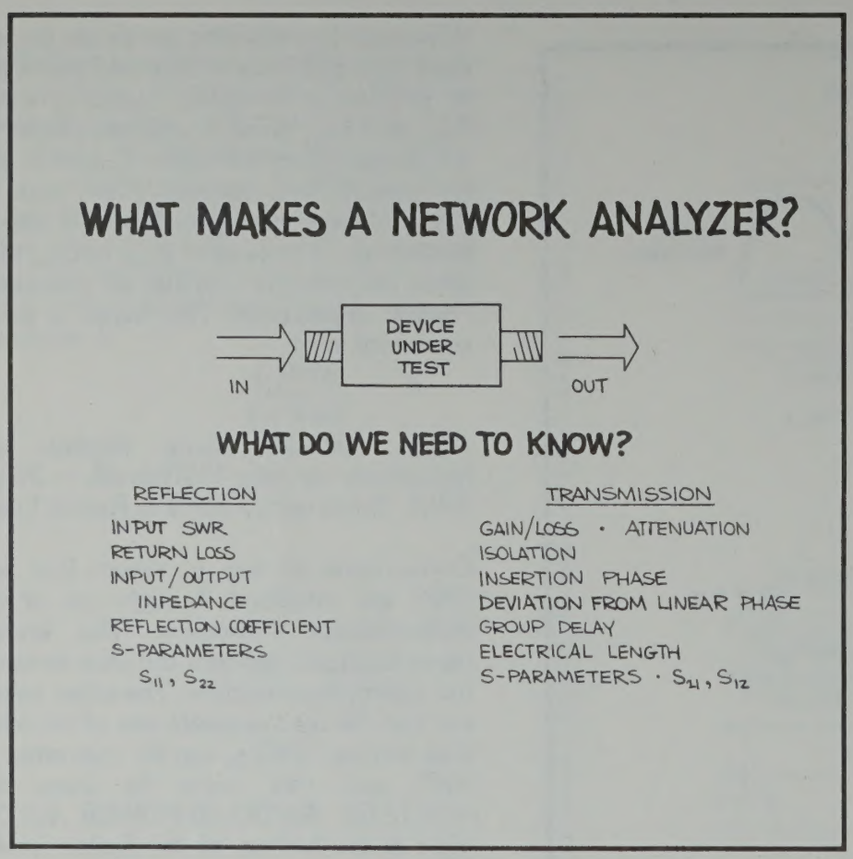
1) A linear change in the input results in a linear change in the output, and

2) The output resulting from multiple input signals is the same as the sum of the outputs resulting from individual input signals.

Simplifying, it can be said that a sine wave input will result in a sine wave output at the same frequency with only an amplitude and phase change. Examples of linear networks are filters, amplifiers, isolators, etc.

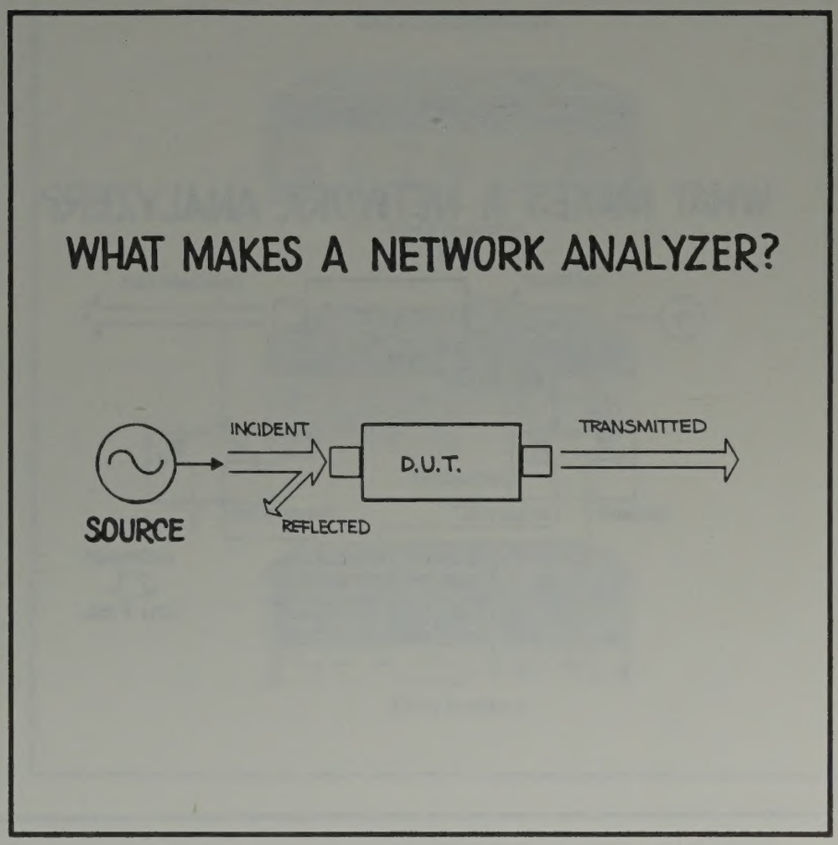
The outputs of nonlinear networks are dependent on the level of the input signal as well as the frequency and are usually composed of multiple frequencies. Examples of such level dependent networks are mixers, diodes, compressed amplifiers, etc. Note the same network may exhibit both linear and nonlinear behavior, such as an amplifier.

Network Analyzers are primarily concerned with the measurement of a networks linear behavior, however, techniques for measuring nonlinear behavior will be discussed where applicable.

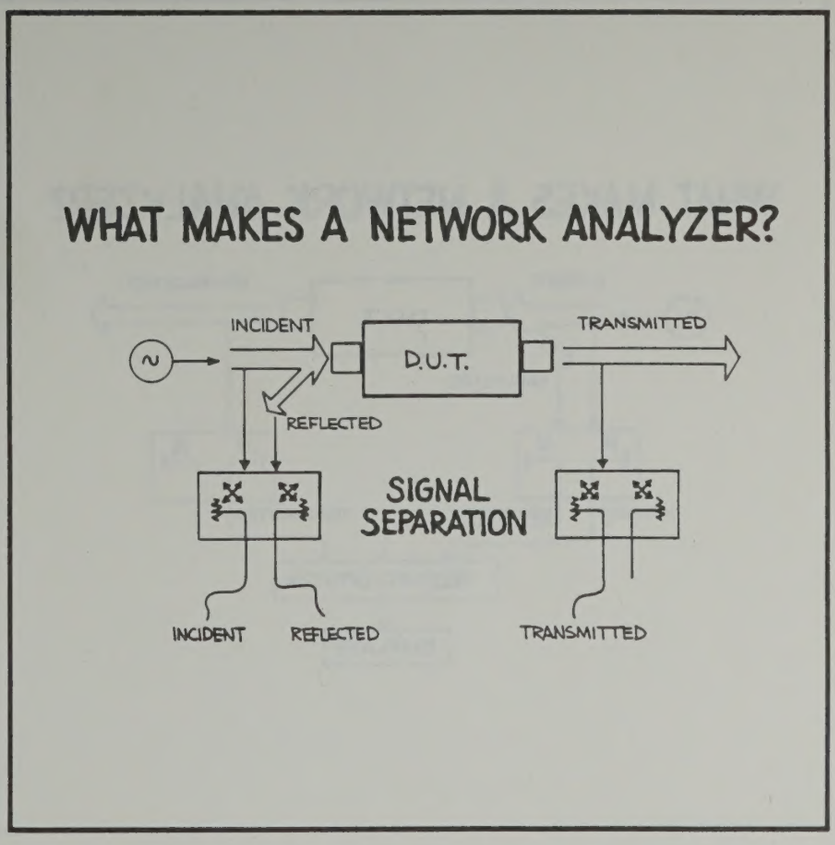


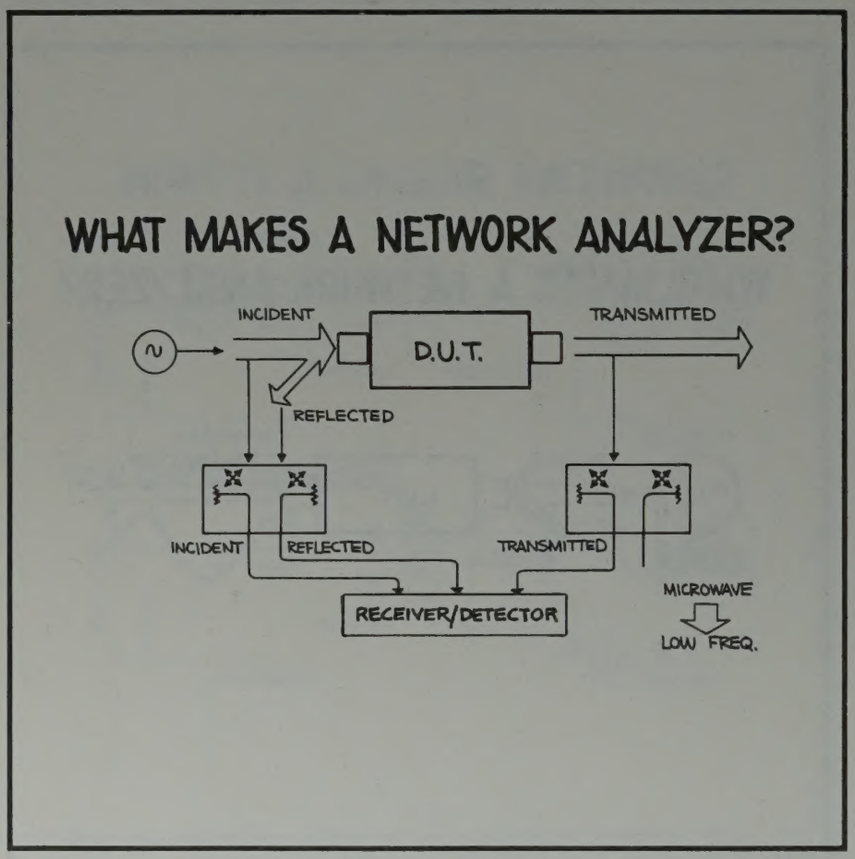
Given a test device, what do we need to know? What test equipment and test equipment configurations are required to measure the transmission/reflection parameters listed on the slide? For the most generality, lets consider a two port test device that is a device with an input and an output (two connectors) and build a measurement system around it.

First, we need a source of RF or microwave energy to produce the incident signal. Once stimulated by the incident signal, the test device responds by dividing the incident into transmitted and reflected signals. By sweeping the frequency of the source the frequency response of the test device can be determined.

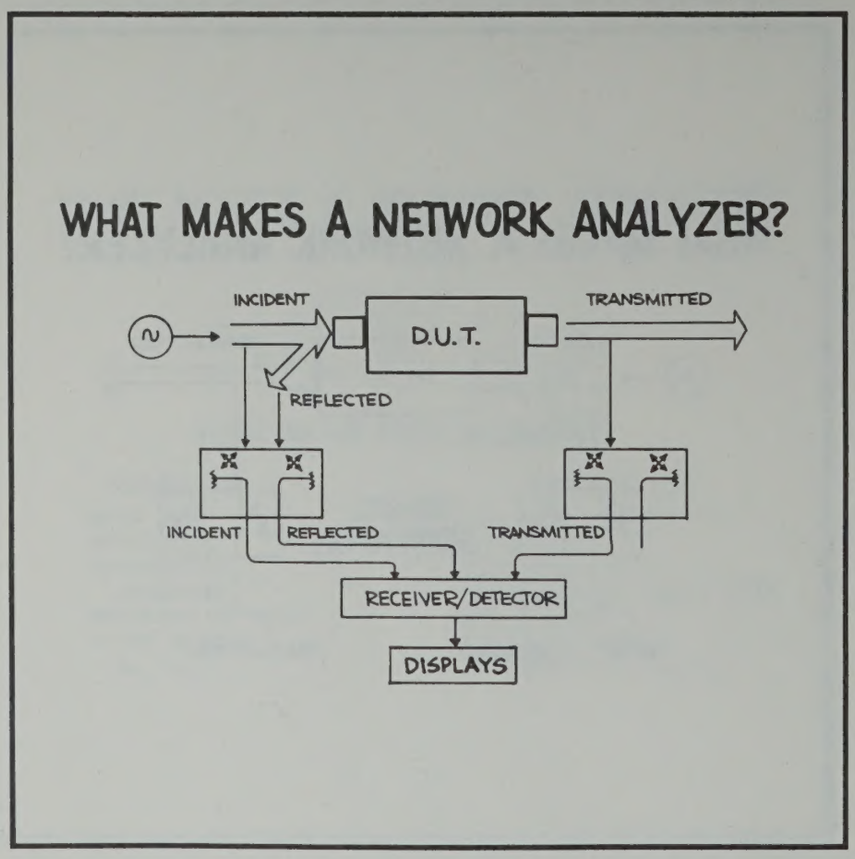


The next step is to sample the incident, reflected, and transmitted signals so that their individual magnitudes and phase differences can be measured. This can be accomplished by directional couplers or directional bridges or even high impedance probes.





Once the appropriate signals have been separated, they must be measured. This is accomplished by a detector or receiver which converts the RF or microwave energy to a low frequency ($\leq$1 MHz) where the signal levels and phase differences can be measured directly. At these lower frequencies, amplitude is essentially measured as the RMS value of a voltage wave. Examples are diode detection systems or mixer/IF receivers.



Finally, the measured data must be displayed in a format appropriate to the information required on the test device. The display may be a swept CRT, a high resolution digital readout, or a computer plot or printout, to name a few.

At this point it is important to note that a Network Analyzer is really made up of four components:

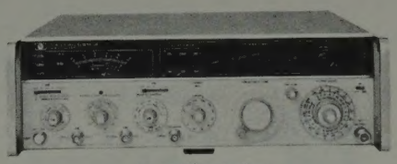
1) A source.
2) A signal separation device.
3) A receiver.
4) A display.

How these components are configured depends on your test device and the data required about the device. Let's briefly review some of the considerations required in selection of a network analyzer system.

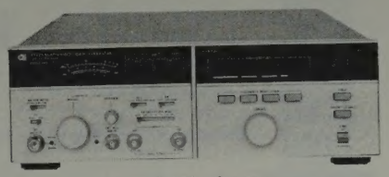
Looking first at sources, there are three basic types of general purpose instruments in this category. Synthesizers and signal generators are primarily used for receiver testing, and as a stimulus for signal analysis. Since they are essentially CW machines, using them for scalar measurements tends to become slow and tedious.

The sweeper, or sweep oscillator, was designed specifically for network measurements. There are several characteristics which are necessary or extremely useful for making these measurements. First, of course, the sweeper must be able to cover the frequency range of interest. A plug-in system provides this capability at the lowest cost to the user. Second, it is desirable to have flexibility in setting the frequency end points of the sweep as well as CW frequencies. To identify specific frequencies (e.g., 3 dB bandwidth) while sweeping, a calibrated marker system is desirable. Some amplitude control and modulation capability is usually necessary. As we'll see later, internal and/or external leveling is almost mandatory. And as automatic systems become more and more cost-effective, programmability becomes more desirable.

SIGNAL SOURCES

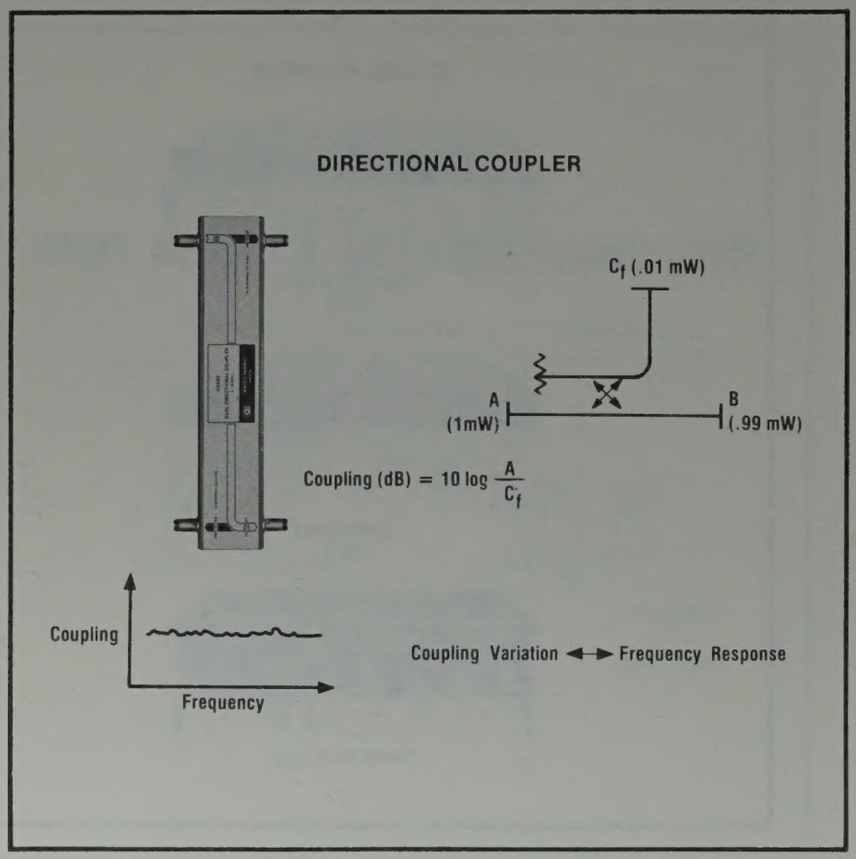


Signal Generators
8640



Synthesizers
8672

Sweep Oscillators



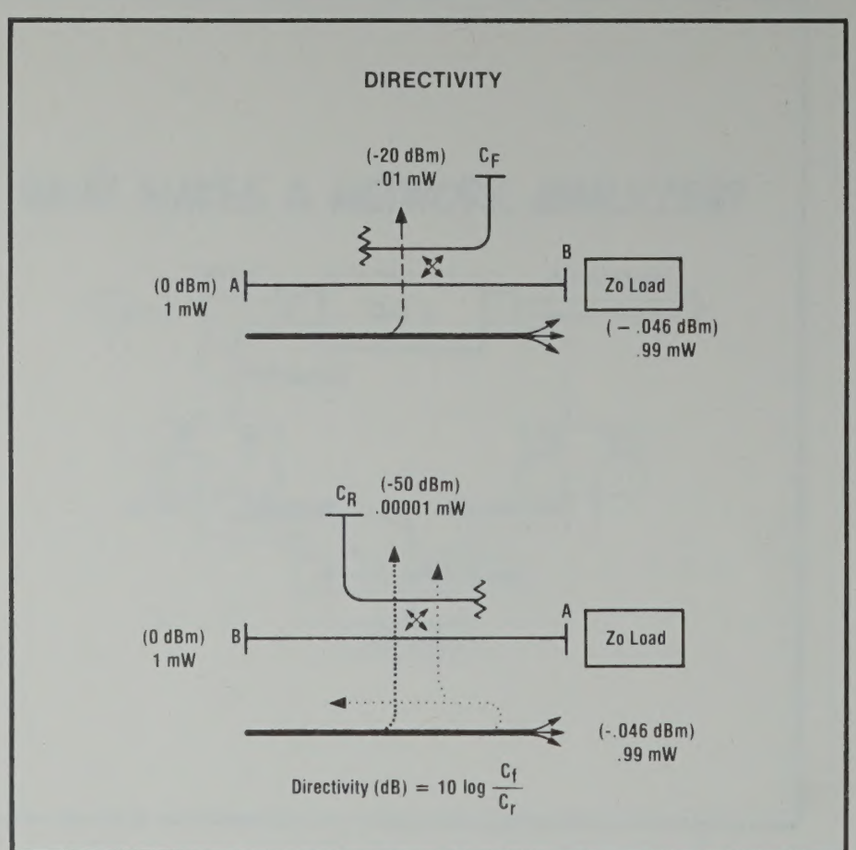
The next component in the reflectometer set-up is the signal separation device. The most popular such device is the directional coupler. It allows us to sample energy traveling in one direction, thus permitting a measure of reflected energy. The coupling factor is a ratio of the power into the coupler to the power out the auxiliary arm. 1 mW into a coupler with 20 dB coupling factor would produce .01 mW at the auxiliary arm. Of course, this would reduce the *main arm* transmission by .01 mW. The main arm coupling loss is then

$$-10 \log \left(1 - \frac{C_f}{A}\right).$$

Use the reflectometer slide rule: Return loss/Mismatch loss scales.

There is also a frequency response or coupling variation associated with couplers usually expressed in ± dB.

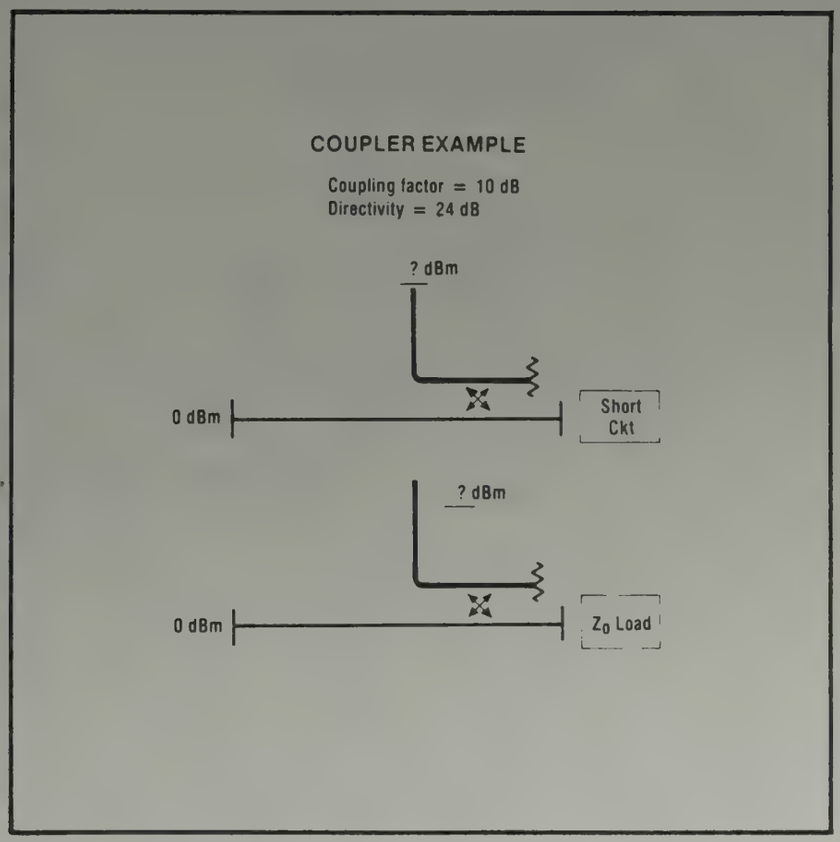
The coupler schematic represents the directions that signals will be coupled depending on which way they are traveling. For example, a signal traveling in the reverse direction (i.e., from B to A) will be coupled into the coupler load. A signal inserted into C would travel toward the coupling load and couple .01 mW toward the A port.



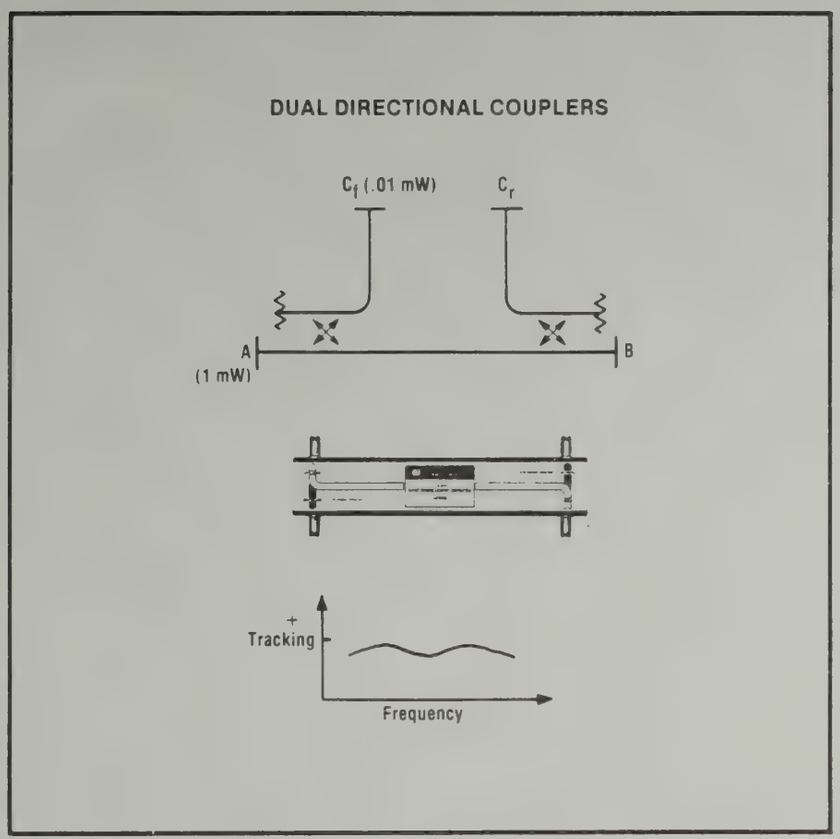
As we'll see later, the most important characteristic for couplers is their directivity. Ideally, signals traveling in the coupler's reverse direction will contribute no signal to the auxiliary arm. In reality, however, some energy does "sneak in the back door". Directivity is the ratio of power coupled into the coupled arm when the coupler is in the forward direction to the power available in the coupled arm when the coupler is in the reverse direction, using a Z_O load in both cases.

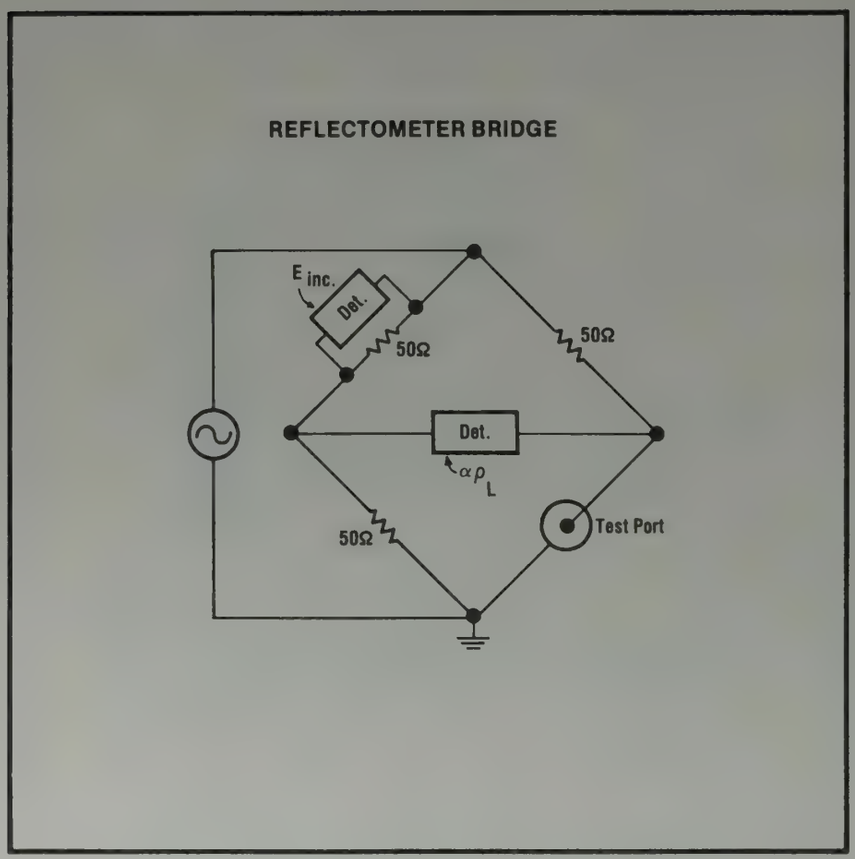
The sources of imperfect directivity are (1) leakage (2) connector reflections and (3) coupler load reflections. Directivity *IS NOT* the same as isolation. Isolation = Directivity + coupling factor: a signal traveling in the reverse direction first encounters the coupling factor (-20 dB) and then is further decreased by the directivity (-30 dB).

For a coupler with these specs, what is the power level appearing at the coupled port in each situation?



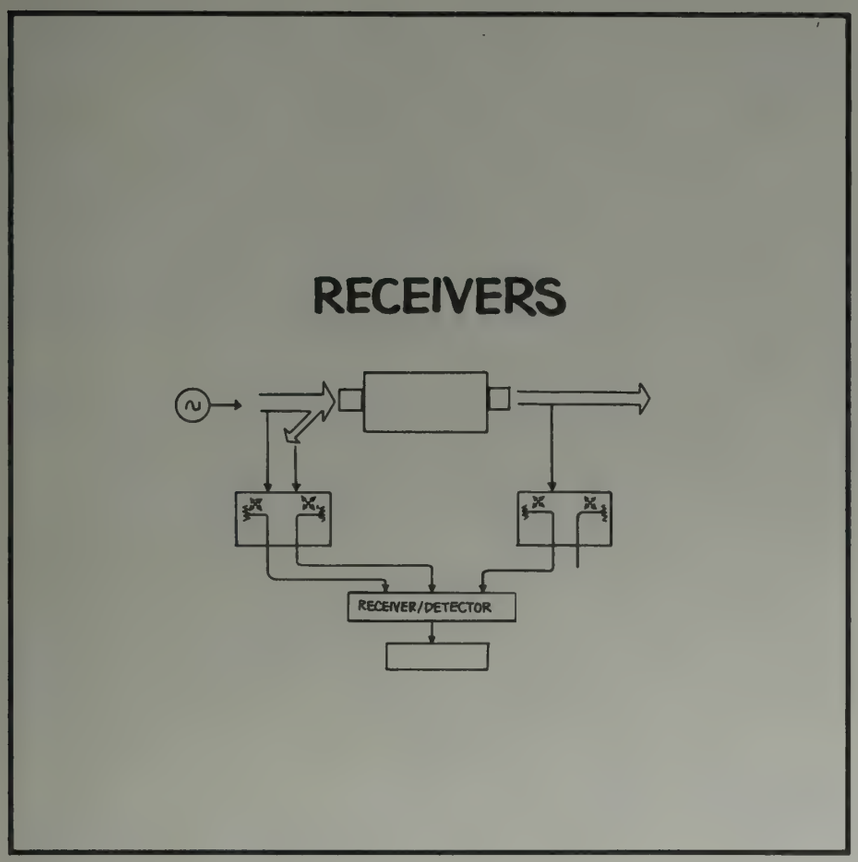
To permit simultaneous monitoring of both incident and reflected signals, two couplers are required. Dual directional couplers provide this capability in one package. Since we will generally be ratioing the two signals (E_{inc}/E_{refl}) in some way, the coupling variation of one arm is virtually canceled by any common variation of the other arm. This results in less system frequency response uncertainty. The figure of merit for how well the arms cancel their frequency responses is tracking.



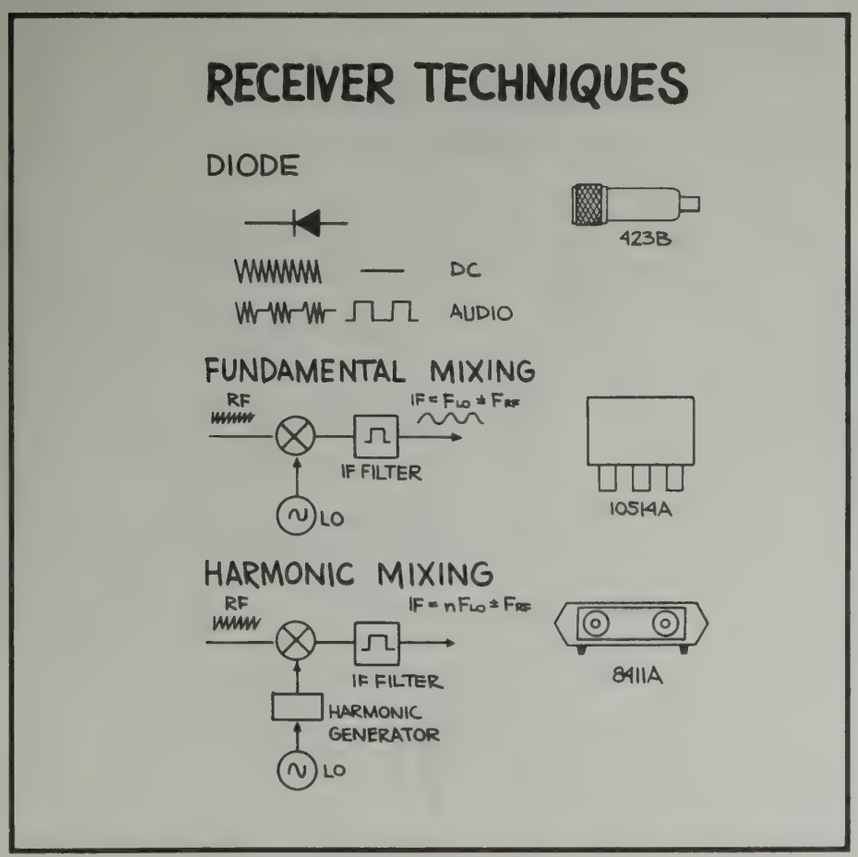


Another signal separation device is the reflectometer bridge. It is similar to the simple Wheatstone bridge. If all four arms are equal in resistance (i.e., Test port = 50Ω) a voltage null is measured. If the Test Port load ≠ 50Ω the voltage across the bridge is proportional to the reflection coefficient of the DUT.

With the other detector sampling $E_{incident}$, a ratio measurement could be made, the benefits of which will be discussed later.

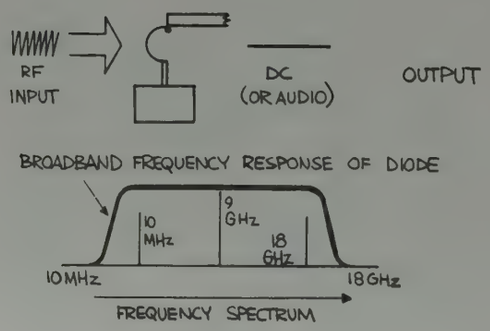


Since it is difficult to measure voltages directly at RF or microwave frequencies, a means of converting to DC or low frequency is required. This is the function of the receiver.



There are three basic receiver techniques used in network analysis. The diode is a broadband device that converts the RF signal level to a proportional DC level. If the signal is amplitude modulated, the diode strips the modulation. Fundamental and harmonic mixing use a mixer and local oscillator or local oscillator harmonic to downconvert the RF to a constant IF frequency which is selected by the IF filter. They act as tuned, low noise receivers, where the frequency tuning and selectivity are determined by L.O. and the IF filter.

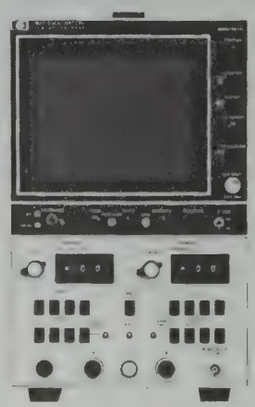
DIODE DETECTION



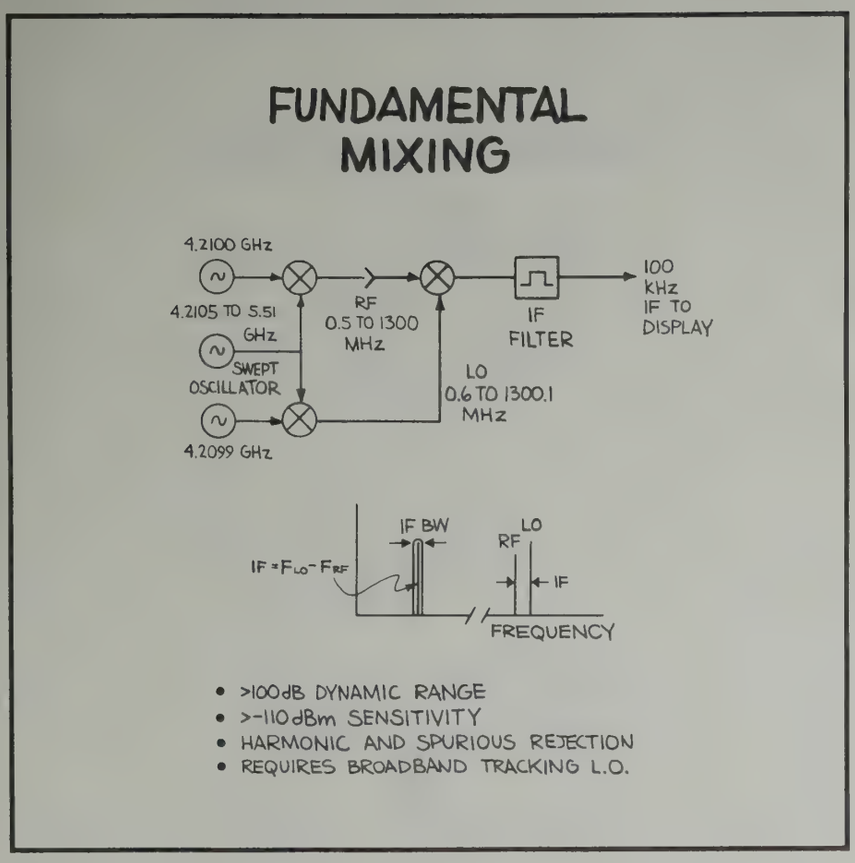
- WIDE FREQUENCY RANGE
- ECONOMICAL
- FAST RESPONSE
- -50dBm SENSITIVITY
- BROADBAND DETECTION

Because of their broad frequency coverag (<1 MHz to >40 GHz), economical cost, and fast response time, diode detectors are widely used for swept frequency measurements with real time CRT displays. Sensitivities of -50 dBm are typical, yielding 60 dB dynamic range when post detection shaping is employed. However, their broadband nature not only limits their sensitivity but also makes them sensitive to the source harmonics, and phase is not measured with most diode detection schemes.

HP 8755 FREQUENCY RESPONSE TEST SET



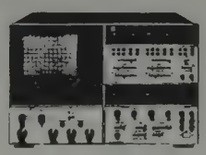
The HP 8755 was designed specifically for swept scalar measurements. It capitalizes on the reliability, sensitivity and repeatability of the Schottky diode. As such, it has a noise floor lower than -50 dBm and is calibrated up to +10 dBm. Three Schottky detectors are employed allowing simultaneous ratio measurements of return loss and transmission gain/loss or of any two interacting or independent parameters. Any of the detectors can also measure absolute power. The detectors are totally interchangeable without recalibration. The display is logarithmic with sensitivities of .25 dB/div to 10 dB/div.



In fundamental mixing, the L.O. always tracks the RF signal by a fixed frequency equal to the I.F. This allows a simple mixer conversion to a fixed low frequency I.F. `This technique provides low noise detection since the narrow IF bandwidth is free from harmonic and spurious responses. Sensitivities are typically -110 dB (10 KHz bandwidth) with 100 dB of measurement range. With the addition of a second channel accurate phase measurements can also be accomplished at the IF frequency. Depending on the frequency coverage desired the tracking L.O. may be both costly and difficult to obtain.

The Models 8505A and 8407A are examples of fundamental mixing receivers. Both of these systems have dedicated sources which provide tracking L.O. outputs.

FUNDAMENTAL MIXING
(SOURCE DEPENDENT RECEIVERS)

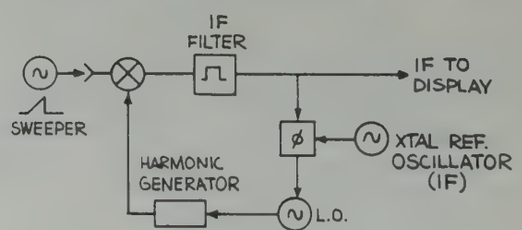


HP 8505A
500KHz-1.3 GHz

HP 8407A
100 KHz-110 MHz

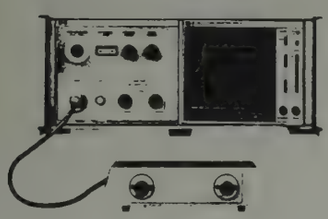
By using the harmonic of a low frequency oscillator as the L.O. signal to the mixer, harmonic mixing receivers obtain extremely wide frequency coverage with a single L.O. For example, the 8410B operates from 100 KHz to 18 GHz with a 60 MHz to 160 MHz L.O. Low noise detection in a narrow IF bandwidth is achieved for harmonic and spurious signal rejection, however, sensitivities and dynamic range are less than those of fundamental mixing because of the added noise produced by the L.O. harmonics. By the addition of a second input channel accurate phase measurements are possible at the IF frequency. Finally, it should be noted that sweep widths are normally limited to octaves because of the limited range of the low frequency L.O. and small variations (<0.1 dB and <1°) can sometimes be noted when different harmonics are used as the L.O. signal.

HARMONIC MIXING



- -80dBm SENSITIVITY
- 60dB TO 80dB DYNAMIC RANGE
- HARMONIC AND SPURIOUS REJECTION
- EXTREMELY WIDE RF COVERAGE
- HARMONIC SKIPPING

HARMONIC MIXING
(SOURCE INDEPENDENT RECEIVER)

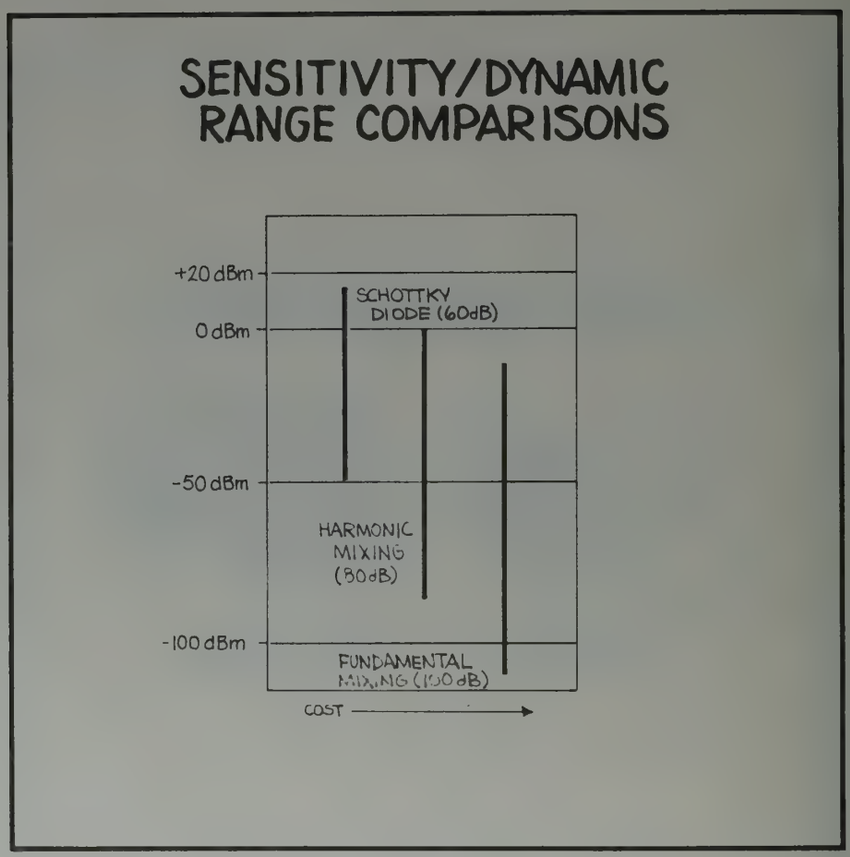


HP 8410B
100 MHz TO 18 GHz

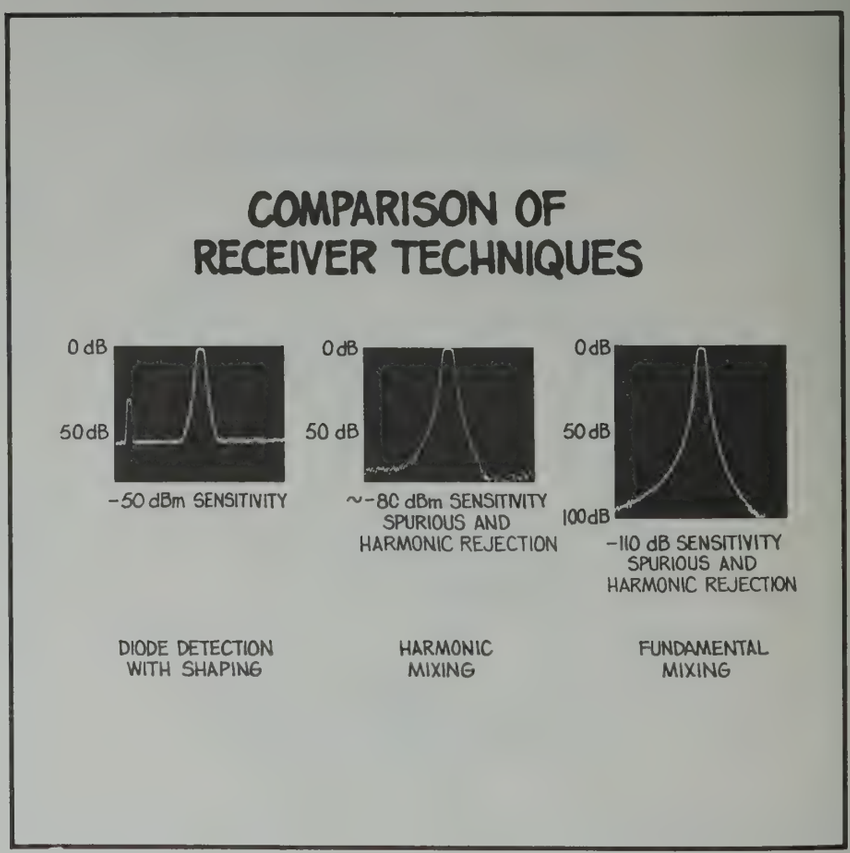
HP 8405B
1 MHz TO 1 GHz

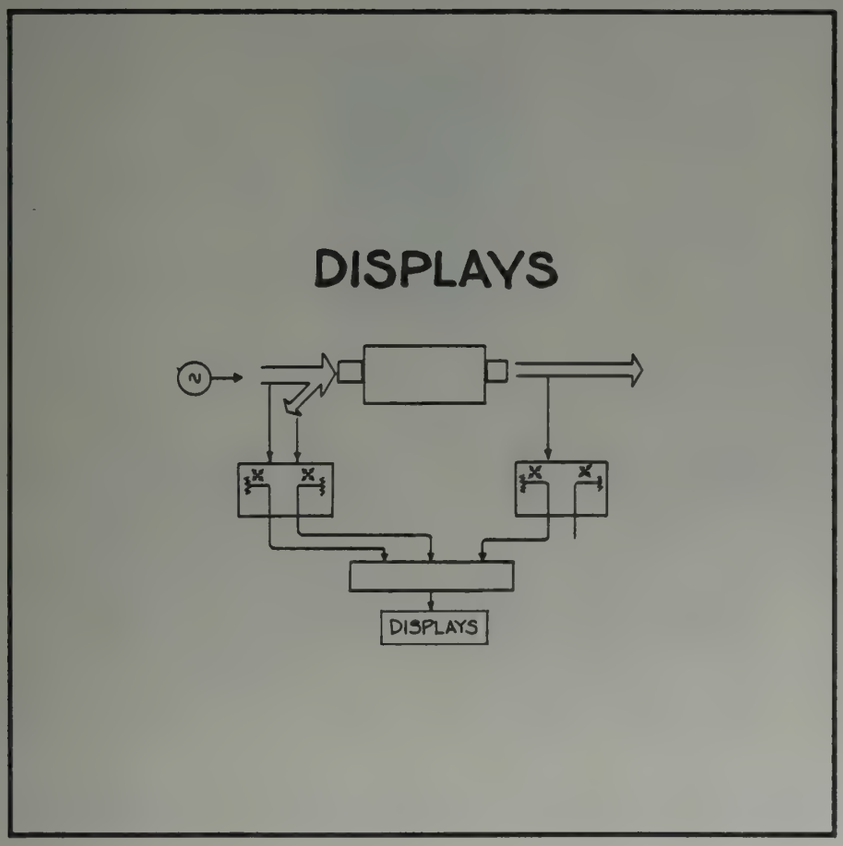
The HP 8410B and 8405A (CW only) are examples of harmonic mixing receivers. They have been designed to operate with virtually any sources, which makes them easily adaptable to a wide variety of applications.

The chart shows the relative dynamic or measurement ranges of the three receiver techniques. Note also the absolute power levels required. For instance +10 dBm is required at the diode for a 60 dB measurement, while only -10 dBm is required for full range with the mixing technique. The sensitivity of the diode detectors (-50 dBm) often precludes the measurement of small signal devices which must be stimulated with low level signals.



A more graphic comparison is obtained when the same bandpass filter is measured using each of the three receiver techniques. Note that harmonic and fundamental mixing provide more dynamic range as well as immunity from annoying harmonic and spurious responses.

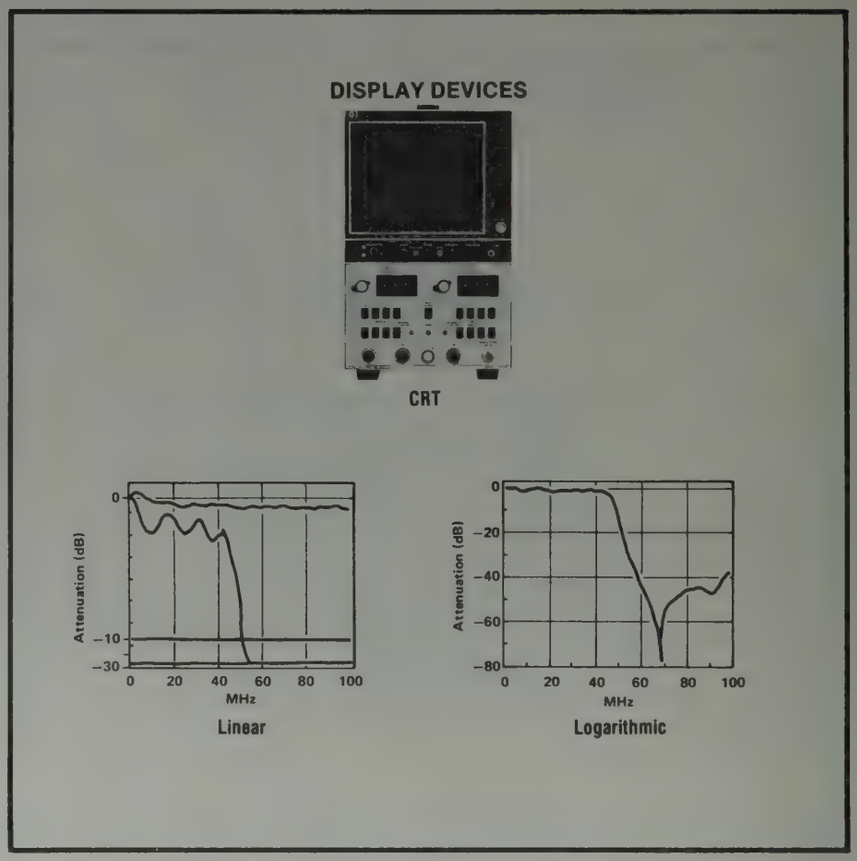




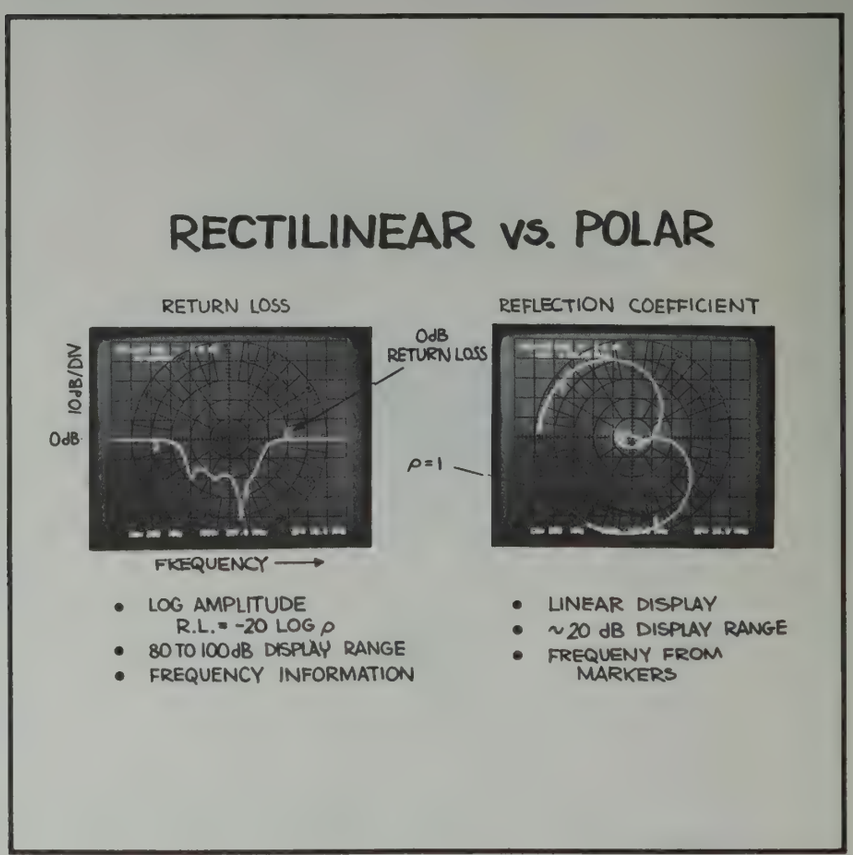
Once the magnitude and phase of the transmission and reflection characteristics have been measured, the data must be displayed in a format appropriate to the information desired about the test device.

There are several ways to display the results in network measurements. Analog and digital meters are often used when high resolution at CW frequencies is required. But for swept real-time measurements, a CRT is the ideal display device. The CRT can be calibrated in either linear or logarithmic units.

Log displays permit very small signals to be observed in the presence of large ones. Linear displays will not show variations in signals which are less than approximately 20 dB below the maximum displayed signal.



The two CRT photos indicate the correspondence between the polar and rectilinear displays. The polar format displays complex reflection coefficient, $\Gamma = \rho\angle\theta$, while the rectangular display shows the same information in a return loss format, R.L. = -20 log ρ. Note the log vs. linear format and the relative dynamic ranges.



This seminar is concerned with scalar (amplitude only) measurements. Phase will be discussed just to the extent that it affects scalar measurement accuracies.

SCALAR MEASUREMENTS

- Magnitude and frequency information only
- Most commonly used measurement technique
- Microwave devices are usually specified in scalar values (SWR, return loss, ρ, gain (dB))
- Scalar measurement systems are less expensive when compared to systems that measure complex parameters (Mag. & Phase).

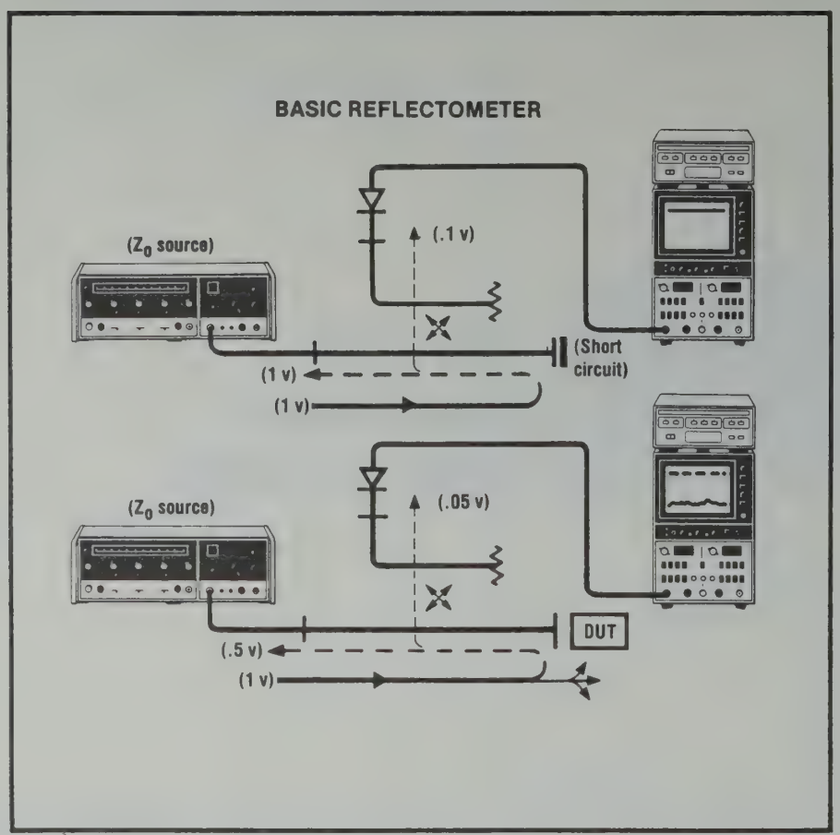
Two methods for making microwave impedance measurements:

1) Reflectometer (most popular)
   a) reflection coefficient
   b) return loss

2) Slotted line
   SWR

First, we need to define these terms...

Scalar Impedance Measurements



There are two parts to the measurement. First, a reference line is established on the CRT with a known standard, in this case a short circuit. This trace is then stored in the 8750A for future subtraction. This establishes a zero dB R.L. line. The short is then replaced with the DUT. The dB change or INPUT-MEMORY is the R.L. of the DUT.

We make a number of assumptions here. (1) The source has a perfect Z_O output impedance. The "source" includes all hardware up to the test port. In other words, the source produces no re-reflections which might cause errors in the measurement. (2) The coupler's directivity is ∞ dB, sampling none of the energy flowing toward the DUT.

Since these assumptions would rarely, if ever, be true, we need to be able to qualify our system to determine if it is accurate enough for the requirements.

This equation is a simplification of a complex flowgraph analysis of reflectometer uncertainties. $\Delta\rho$ is the worst case uncertainty in the measurement where ρ_L is the measured reflection coefficient of the DUT. A, B, and C are all in linear terms; i.e., 40 dB directivity converts to A = 0.01. The Reflectometer Calculator can be used to make this conversion: place dB directivity under blue arrow across from R.L. and read linear value across from ρ. Each term in this equation will be analyzed separately.

IMPEDANCE MEASUREMENT ACCURACY

$$\Delta\rho = A + B\rho_L + C\rho_L^2$$

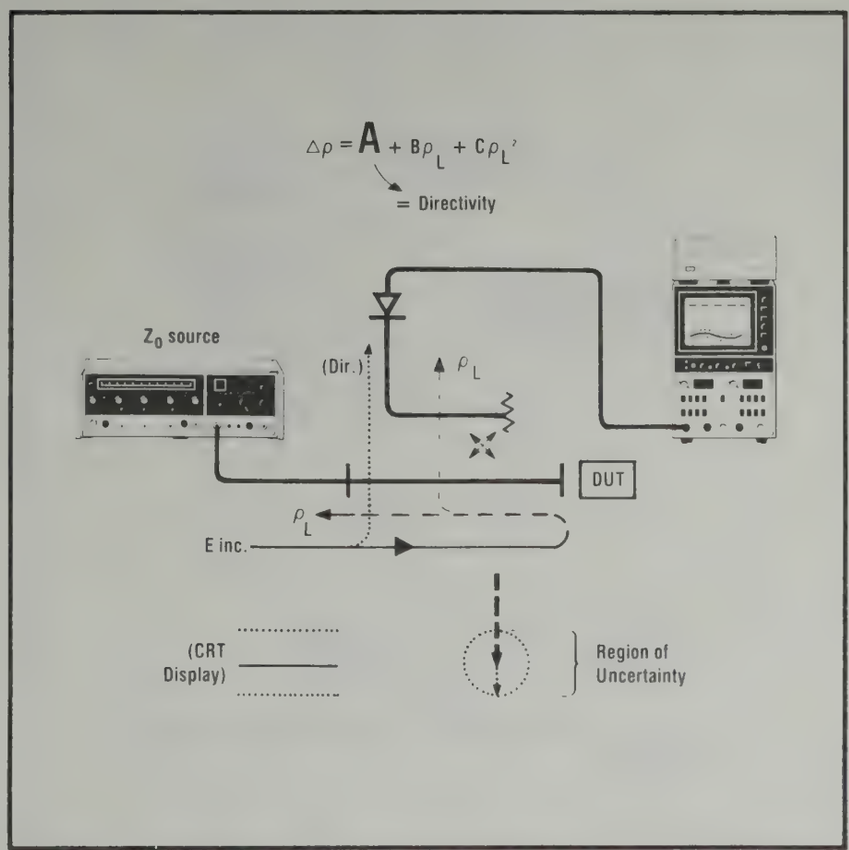
A = Directivity (ratio)

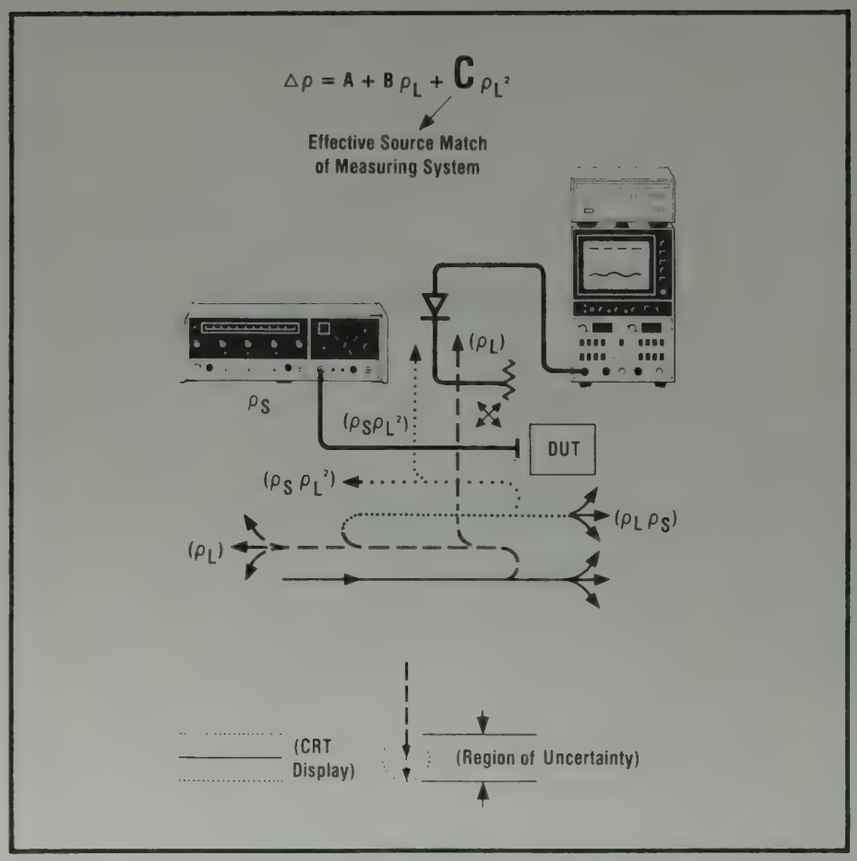
B = Calibration error, frequency response, display and instrument errors

C = Effective source match

Directivity produces uncertainty in all measurements regardless of ρ_L. ρ_L is the value we want to measure but the signal present due to directivity will add at some phase to the ρ_L vector.

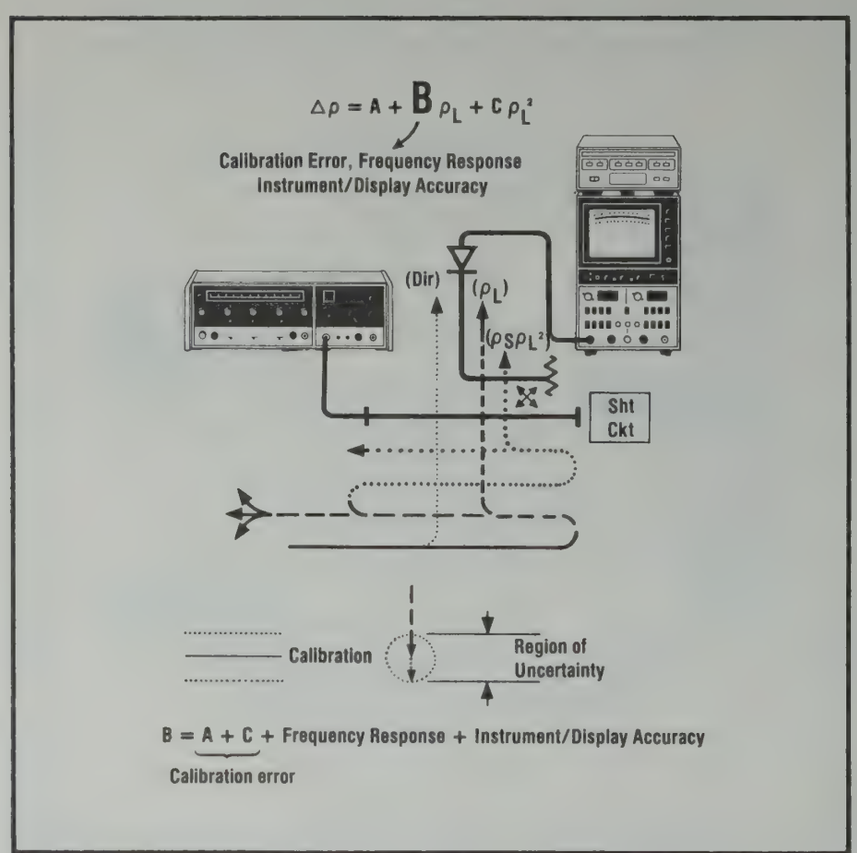
Since we have no phase information in scalar systems, we assume a worst case situation where the directivity vector adds in phase and 180° out of phase with the ρ_L vector. This gives us an uncertainty window about the measured value due to imperfect directivity. Since it is always present regardless of ρ_L, directivity is the limiting factor in measurements of DUT's with a low ρ_L.





By following the flowgraph, it can be seen that effective source match belongs in the C term. The first reflection from the DUT is ρ_L; this is what we want to measure. This signal continues back toward the source where it is re-reflected if ρ_S is not perfect, resulting in a signal normalized to $\rho_S \times \rho_L$ flowing back toward the DUT where it is again reflected and sampled as $\rho_S\rho_L^2 = C\rho_L^2$.

Of course, we are assuming directivity = ∞. As with the A term, this $\rho_S\rho_L^2$ term could be at any phase; we assume worst case conditions. Source match then results in a worst case uncertainty of $\pm\rho_S\rho_L^2$; the larger ρ_L, the more important source match becomes.



The A and C terms were concerned with errors in the test portion of the measurement. It makes sense that the same errors are present in calibration. Directivity, of course, is present regardless of the load. Since $\rho_L = 1$ when a short circuit is attached $\rho_S\rho_L^2 = \rho_S$.

Since the calibration stage essentially sets the gain of the test system, a $\pm 10\%$ error in calibration results in a $\pm 10\%$ error in the measured value. This is the reason the calibration error is in the B term (i.e., multiplied by ρ_L). Similarly, frequency response and instrument accuracy are scaled by ρ_L; so they are also part of B term. Primary emphasis here, however, is on B = A + C because this uncertainty is present regardless of the instrumentation used.

Now we are prepared to evaluate our test set-up. Using typical values for directivity and source SWR and assuming a measured ρ_L of 0.5, we come up with typical worst case errors of ±.28 or ±56%! In other words, the SWR of the DUT is somewhere between 1.56 and 8.2:1.

Let's see how we might improve the uncertainty in this measurement. We obviously want to make A, B, and C coefficients as small as possible.

HOW ACCURATE WAS OUR REFLECTOMETER MEASUREMENT?

$$\Delta\rho = A + B\rho_L + C\rho_L^2$$

Directivity 30 dB (.0316)

Source SWR 1.9:1 (0.31)

A = .0316
B = 0.3416
C = 0.31

} Without Instrument Error

With ρ_L measured at 0.5

$$\Delta\rho = .0316 + .3416 \times .5 + .31 \times .5^2$$

$$\Delta\rho = \pm 0.28$$

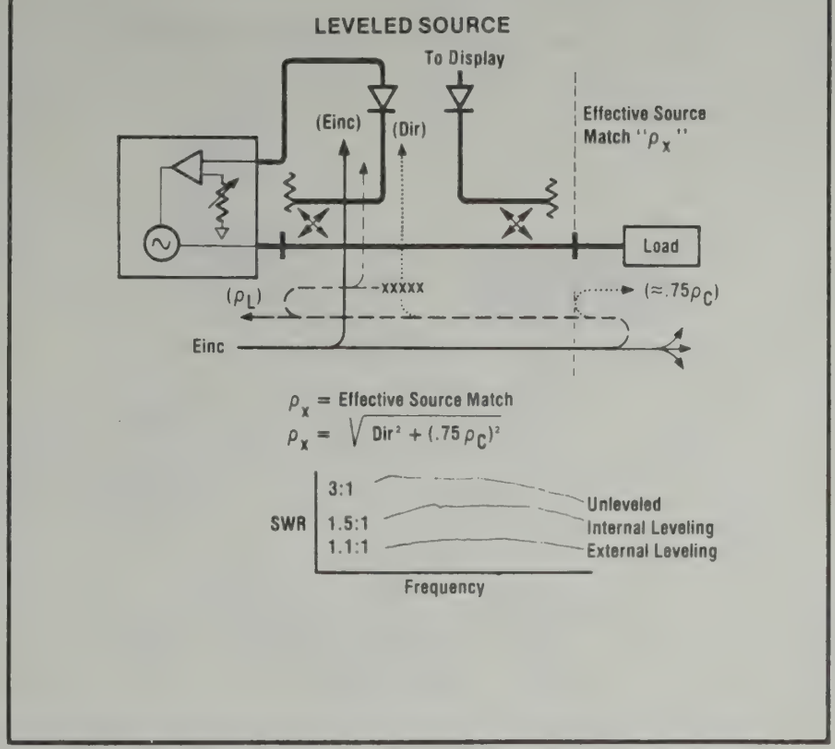
Improving effective source match can be simple and inexpensive. Leveling the sweeper output does this by varying the power to the load to counteract source reflections. Any signal re-reflected from the sweeper is sensed by the leveling loop which corrects the output from the sweeper accordingly.

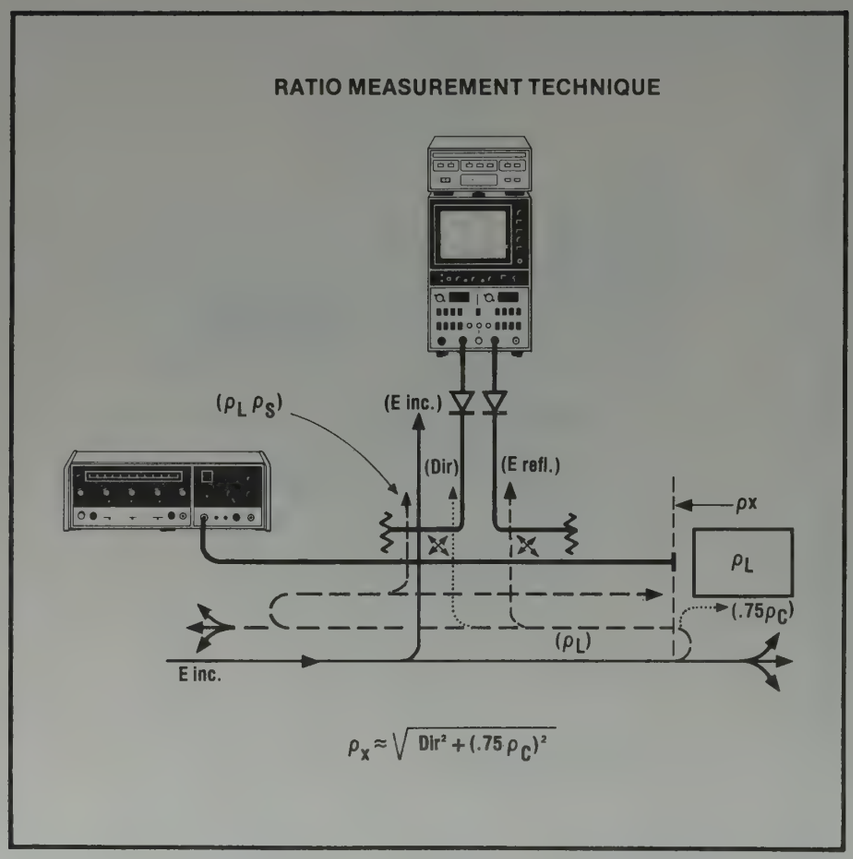
Although leveling noticeably improves source match, there are still inherent uncertainties. First, the output connector of the leveling coupler has some reflection. In addition, that same old directivity gets us here. The equation gives a good approximation of the new effective source match, ρ_X.

ρ_C as defined on most couplers is the mainline mismatch which includes both the input and output connectors. Since the input connector is within the leveling loop, its effects are compensated; therefore, we scale ρ_C by .75. The equation very closely approximates the detailed flowgraph analysis of a leveled source.

Since ρ_X is defined as being at the measurement plane, the mismatch introduced by the reflection-sampling coupler must be accounted for. With a dual directional coupler, ρ_C accounts for both couplers' mismatch. When two separate couplers are used, however the effects of the connectors of both must be considered.

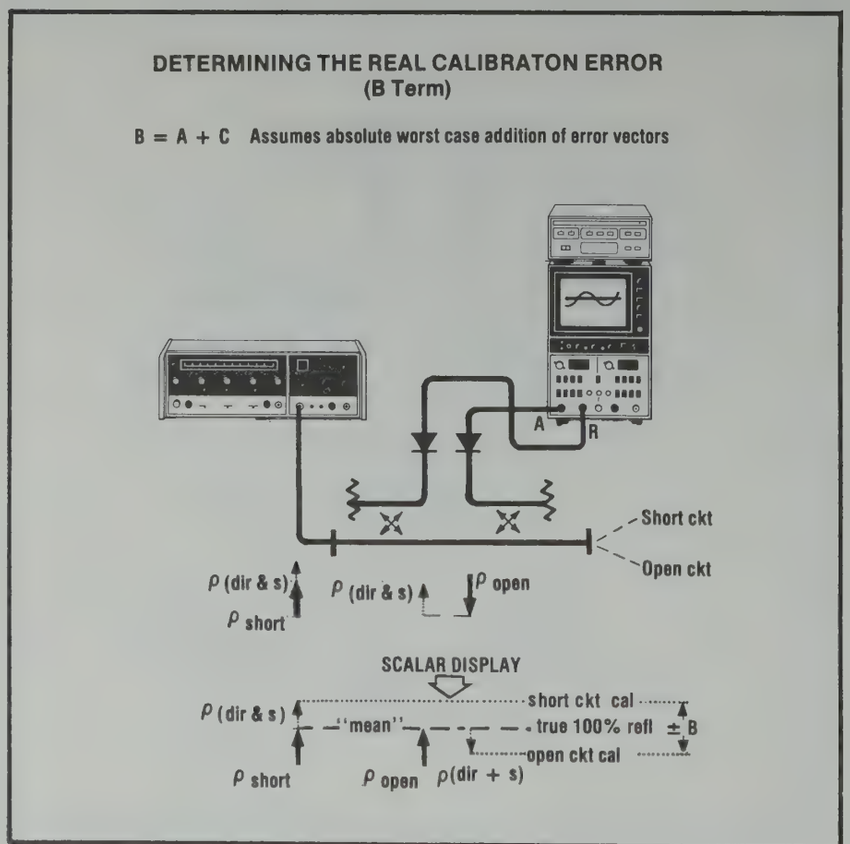
Where possible then, a dual directional coupler is recommended.





Effective source match can also be improved by ratioing the transmitted and reflected signals. In this technique the absolute incident power is not controlled but the variations are ratioed out. Any re-reflections ($\rho_L\rho_S$) are seen by both detectors, and when you take the ratio A/R on the 8755 the effect of $\rho_L\rho_S$ is cancelled. Again, since the dual directional coupler is not perfect, the "effective source match" must be calculated. The effective source match is determined by the same equation as for leveling.

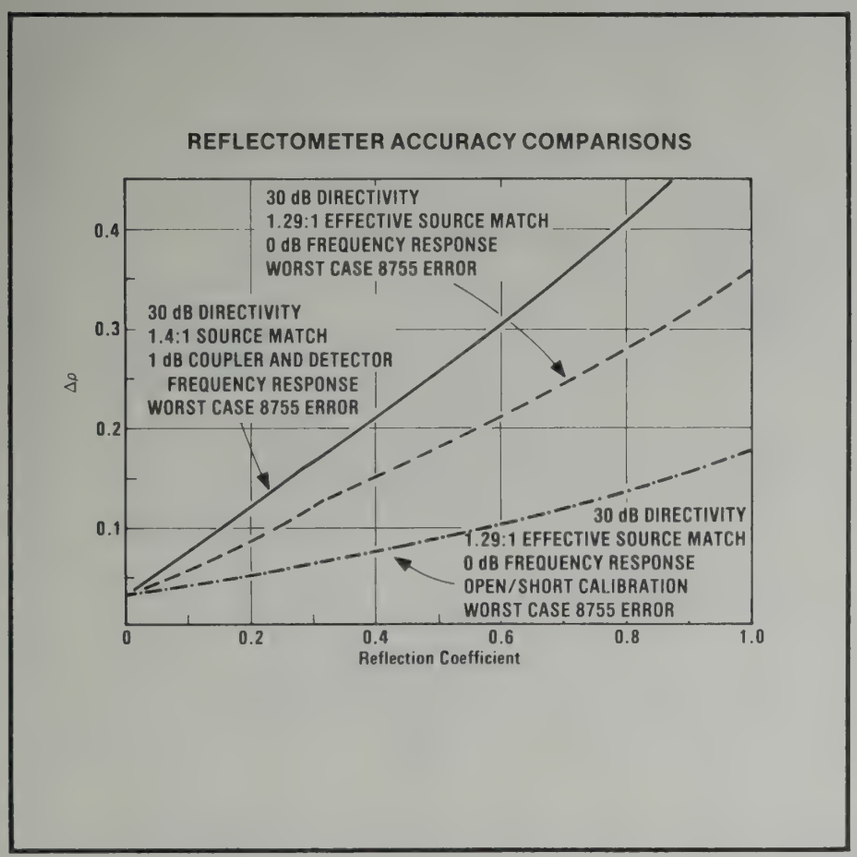
The advantage of ratioing is that the source need not have a leveling capability; i.e., the "source match," is source- independent. In some cases, however, it may not be usable with power level- sensitive DUT's unless internal leveling is available.



The B term is a major contributor to $\Delta\rho$. We improved it partially by minimizing the C term ($B = A + C + \ldots$). This still assumes worst case addition/subtraction of the errors. Since this is virtually impossible it would be advantageous to determine the real value for the B term.

The reflection from an open circuit is 180° out of phase with that from a short circuit, but the errors due to directivity and source match do not change phase when ρ_L is changed from open to short (Directivity is independent of ρ_L; $\rho_s\rho_L^2$ experiences ρ_L twice x 180° = 360° = 0°.)

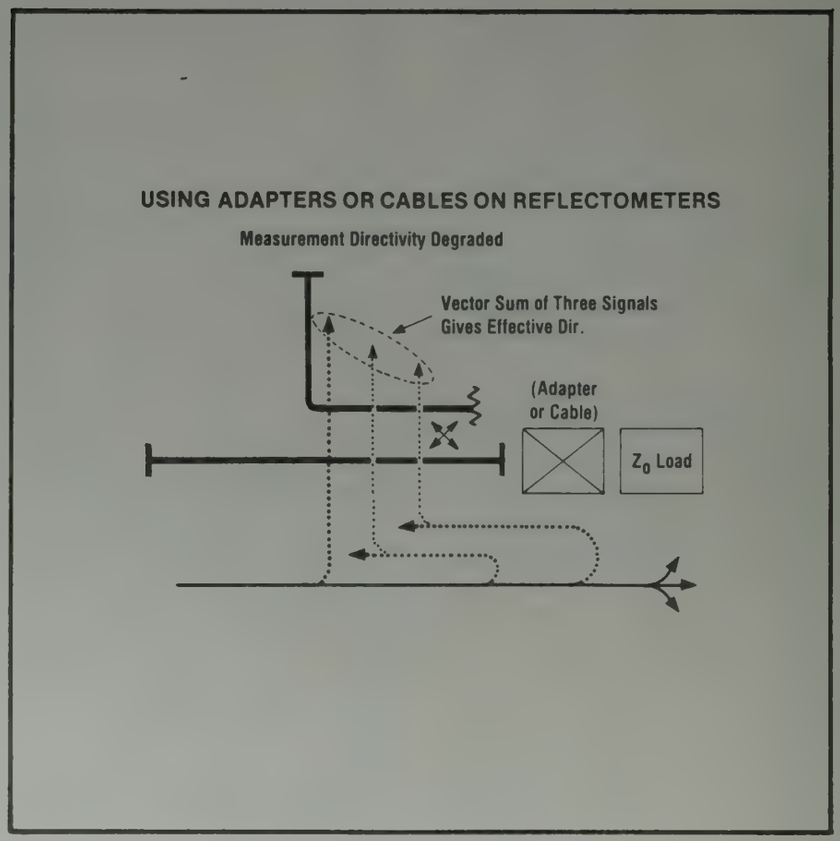
We can use this fact to determine the actual value of B during a "pre-calibration" of the test system. First, a short circuit is attached and the response stored by the 8750A. Then the short is removed and the output port is left open circuited. INPUT - MEMORY then displays B peak-to-peak in dB. To convert to $\pm B_{linear}$ use the reflectometer calculator for a close approximation: dB to Voltage Ratio scale. Then, since this is a peak-to-peak value relative to 1, subtract 1 (to obtain peak-to-peak error) and divide by 2 (for peak error). For example a 2 dB ρ-ρ display is the same as $B \approx 0.13$.



This shows the relative effects on $\Delta\rho$ of the three measurement techniques vs. the measured ρ_L. The top curve is that of the simple reflectometer; the middle curve is that of the ratio technique without open-short pre-calibration; and the bottom one adds the open-short technique and assumed a very small contribution from the B term.

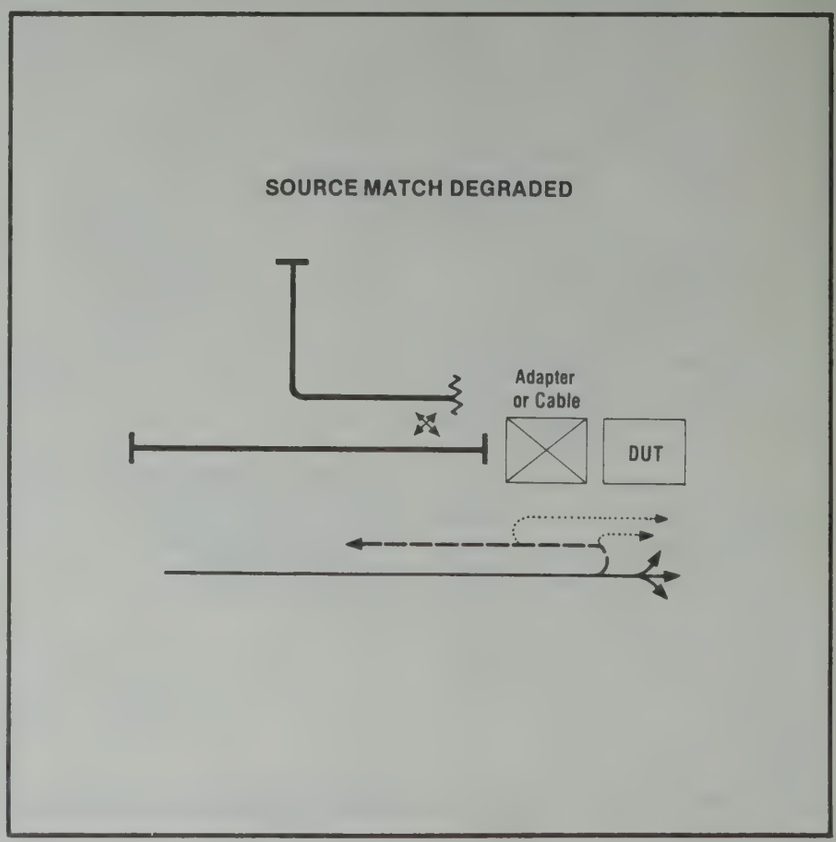
Cables and adapters are often overlooked as contributors to measurement error. However, they can seriously affect the uncertainty, both in the directivity and effective source match.

The effects of adapters on effective directivity are most often misunderstood. As the flow-graph shows, the adapter has the same relationship to directivity as the output connector on the coupler. If the adapter has an SWR of say 1.5 (the $2.00 variety), the effective directivity of the coupler drops to no better than 14 dB, even if the coupler itself has infinite directivity. In other words, with a perfect Z_o load ($\rho_L = 0$) on the output of the adapter; the reflected signal appearing at the coupled arm would only be 14 dB less than the reflection from a short circuit. So, even using all the other information we've learned, the uncertainty window has drastically increased due to the decrease in effective directivity.



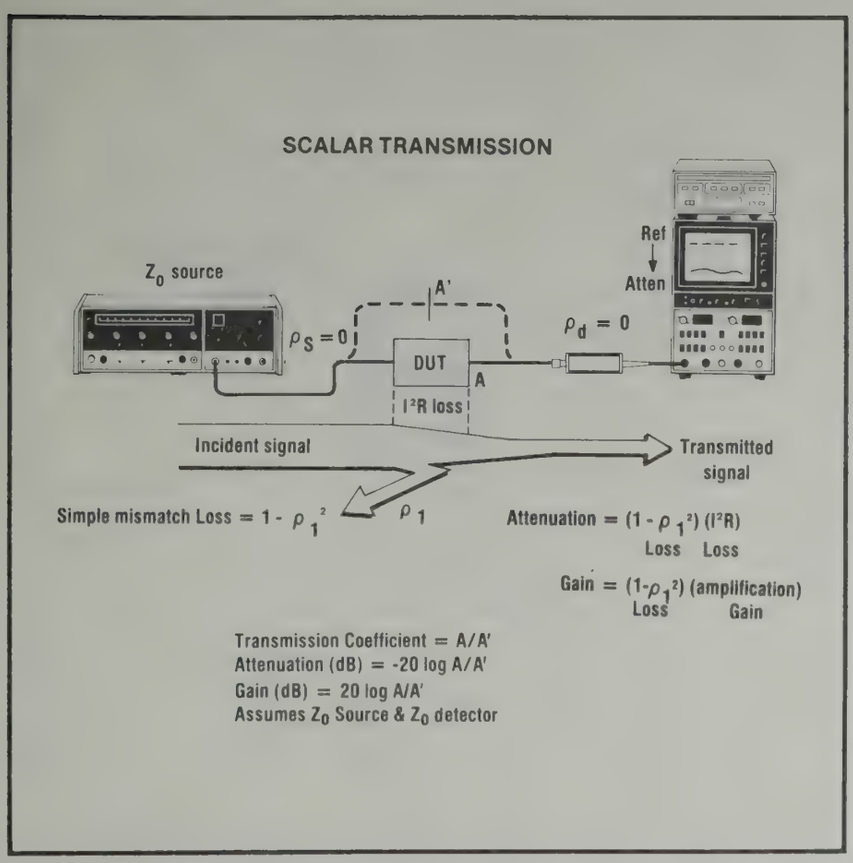
Unfortunately, adapters and cables also degrade effective source match as shown in the flowgraph. The signal reflected off the DUT would normally be re-reflected off the source but then compensated for by either leveling or ratio technique. The adapter now re-reflects a portion of ρ_L so it never gets into the leveling system. This reduces source match to no better than the adapter reflection coefficient.

The lesson for reflectometer measurements is: don't use adapters or cables outside the leveling/ratio loop; or if adapters (never cables!) are necessary use only the best available. In addition, *all* connectors degrade with usage; care, occasional re-calibration, and timely replacement are necessary especially where the measurements are made on DUT's with low ρ_L.



Scalar Transmission Measurements

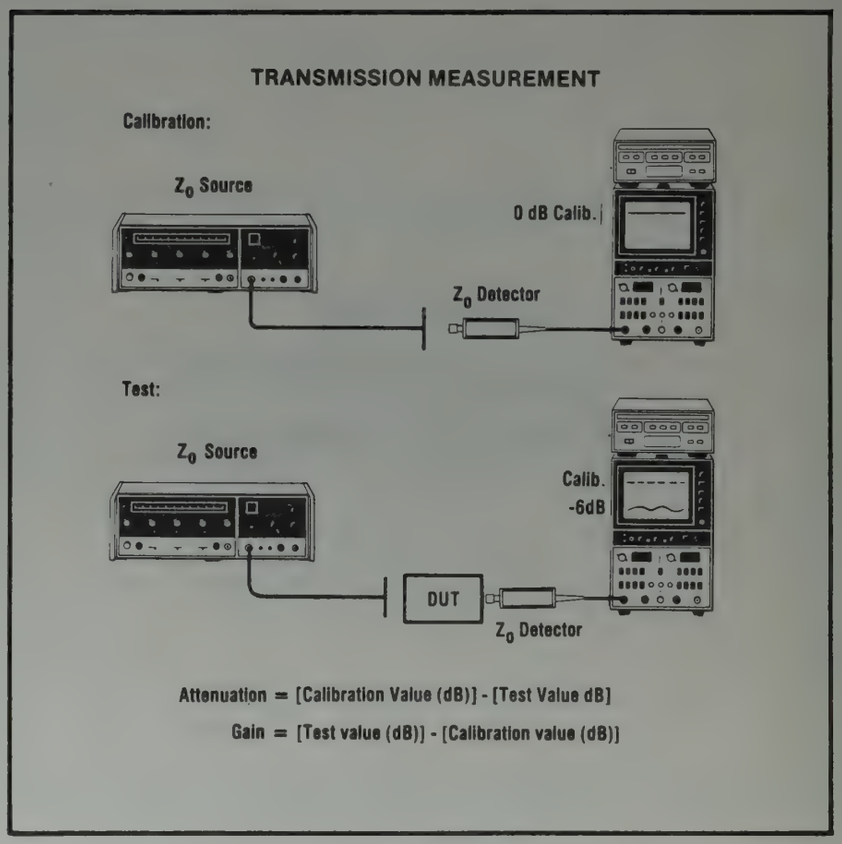
The two basic measurements made to characterize linear networks are impedance and transmission. This section will describe the transmission measurement and investigate the associated errors.



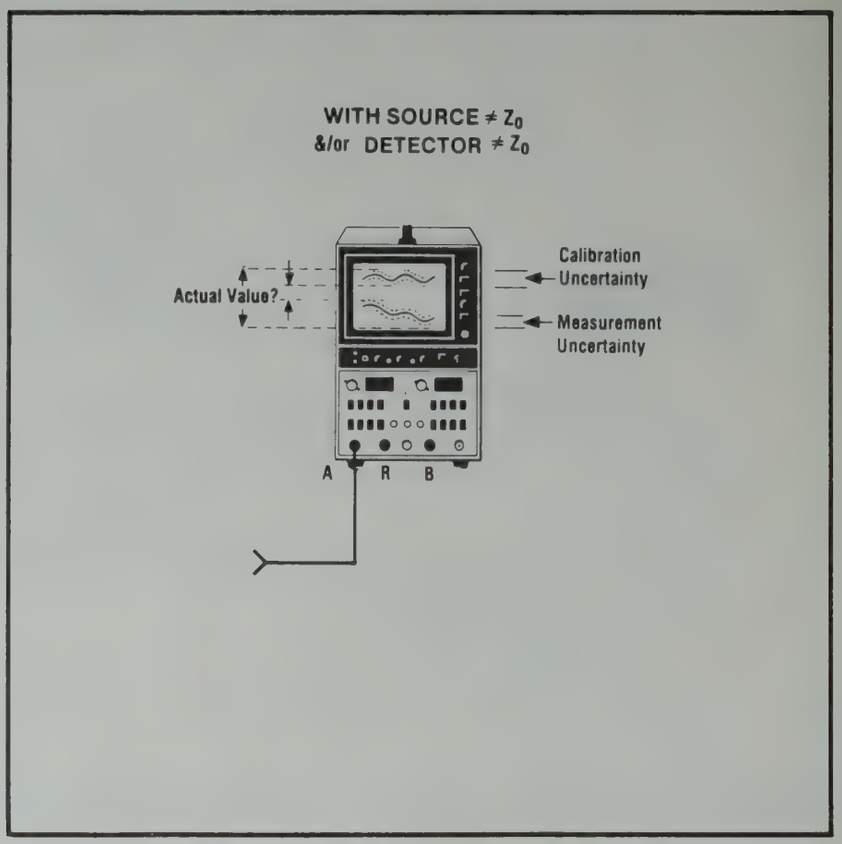
Scalar transmission is defined as the change in power at a detector resulting from the insertion of a DUT between a reflectionless source and a reflectionless detector. Transmission loss is made up of two parts: simple mismatch loss or the energy reflected from the DUT, and I^2R loss, the energy absorbed by the DUT. The mismatch loss is due to the DUT not being Z_O.

Gain is similarly made up of two parts: mismatch loss and internal amplification. As with attenuation, gain measurement accuracy is qualified by source and detector mismatches.

A simple transmission measurement system consists of a source, detector and display. First, the system is calibrated without the DUT (A′); then the DUT is inserted (A) and the difference in dB is the transmission gain/loss of the DUT. To be accurate requires an ideal measurement system. Besides accurate instrumentation, the ideal system is one in which ρ_s and ρ_d equal zero. Since this rarely happens, we need a means for analyzing the system we are using.

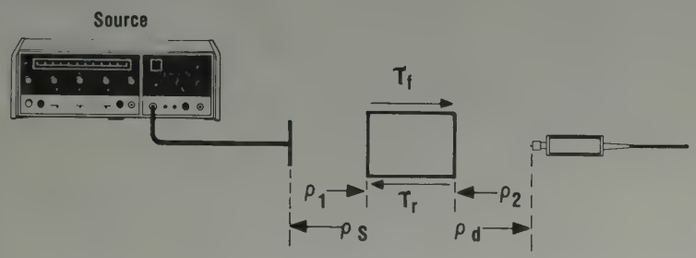


As in impedance measurements, errors are present in the calibration as well as in the measurement phases of the test. These result in an uncertainty window around the two traces. The *actual* value of transmission (loss in this case) can be anywhere between (a) the maximum calibration minus minimum measurement values and (b) the minimum calibration minus maximum measurement values.



TRANSMISSION UNCERTAINTY

$$\text{Transmission Uncertainty} = \frac{\overbrace{(1 \pm \rho_S \rho_d)}^{\text{Calibration Error}}}{\underbrace{(1 \pm \rho_S \rho_1)(1 \pm \rho_d \rho_2) \pm (\rho_S \tau_f^2 \tau_r^2 \rho_d)}_{\text{Measurement Error}}} + \text{Frequency Response} + \text{Display Error}$$

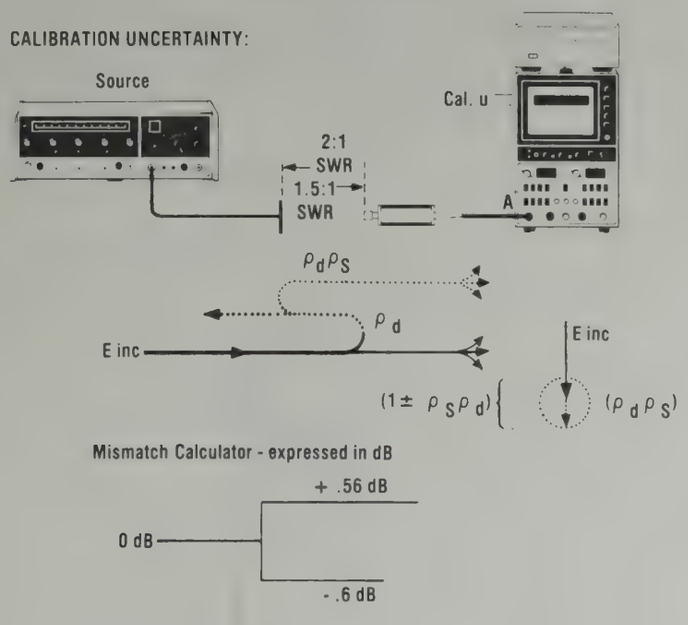


Uncertainties due to impedances not being Z_0 ($\rho_s \neq 0$, $\rho_d \neq 0$).

This transmission uncertainty equation quantifies the worst case error window caused by the system. Frequency response, of course, may be eliminated by the 8750A storage-normalizer. Display errors can be simply added to the uncertainty. The rest of equation applies regardless of the instrumentation. Since it's a tedious exercise to evaluate this linear equation, let's explore an easier and more understandable quantification.

MISMATCH UNCERTAINTIES

CALIBRATION UNCERTAINTY:

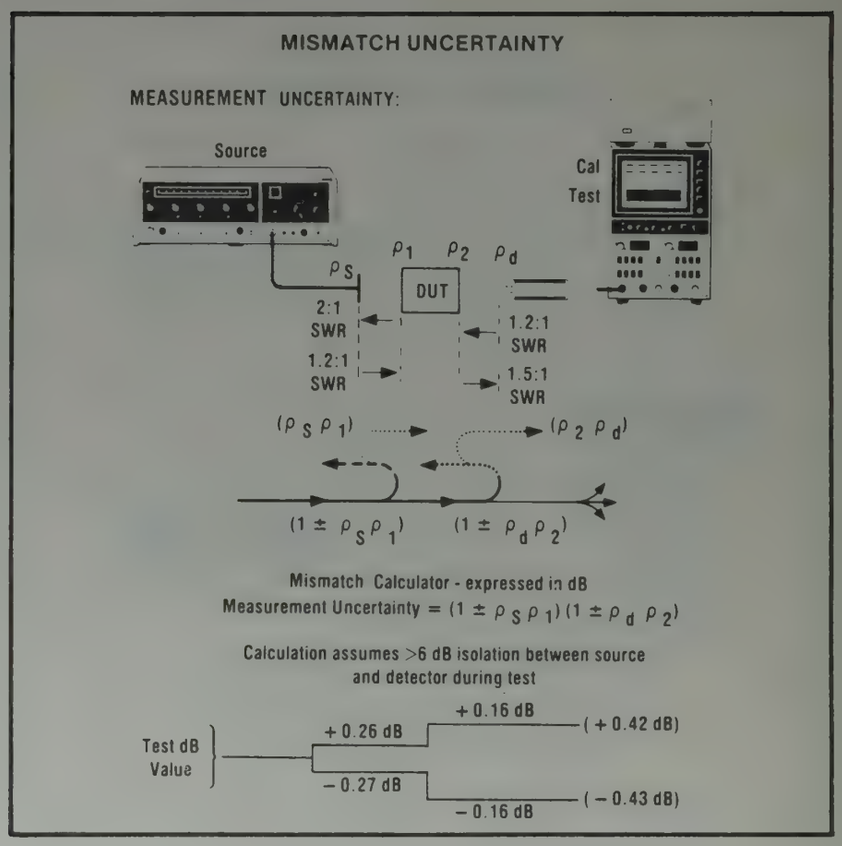


Mismatch Calculator - expressed in dB

First, we'll investigate the uncertainties associated with the calibration stage of the measurement. As can be seen from the diagram, when the detector is attached to the source, the incident signal first encounters the detector mismatch where a signal normalized to ρ_d is reflected. This signal is then re-reflected from the source resulting in an uncertainty vector of $\rho_s \rho_d$ at some unknown phase relationship to E_{inc}. Worst case, the extremes of the signal seen by the detector would be $1 \pm \rho_s \rho_d$; this is the error window in calibration. The Mismatch Error Limits side of the reflectometer calculator converts the two SWR's into the uncertainty limits in dB: Place the black arrow over the SWR of either the source (2 in this example) or the detector (1.5); under the SWR value of the other device read the (+) value of Max Mismatch Error Limits (+ .56 dB); directly below this is the (-) value (-.6 dB).

Similarly, there are uncertainties in the measurement stage due to (1) the source/DUT input mismatch ($1 \pm \rho_S\rho_1$) and (2) DUT output/detector mismatch ($1 \pm \rho_2\rho_d$). Each of these uncertainties can be found with the Mismatch Error Limits Calculator just like before. They are then added to get the uncertainty window in the measurement stage.

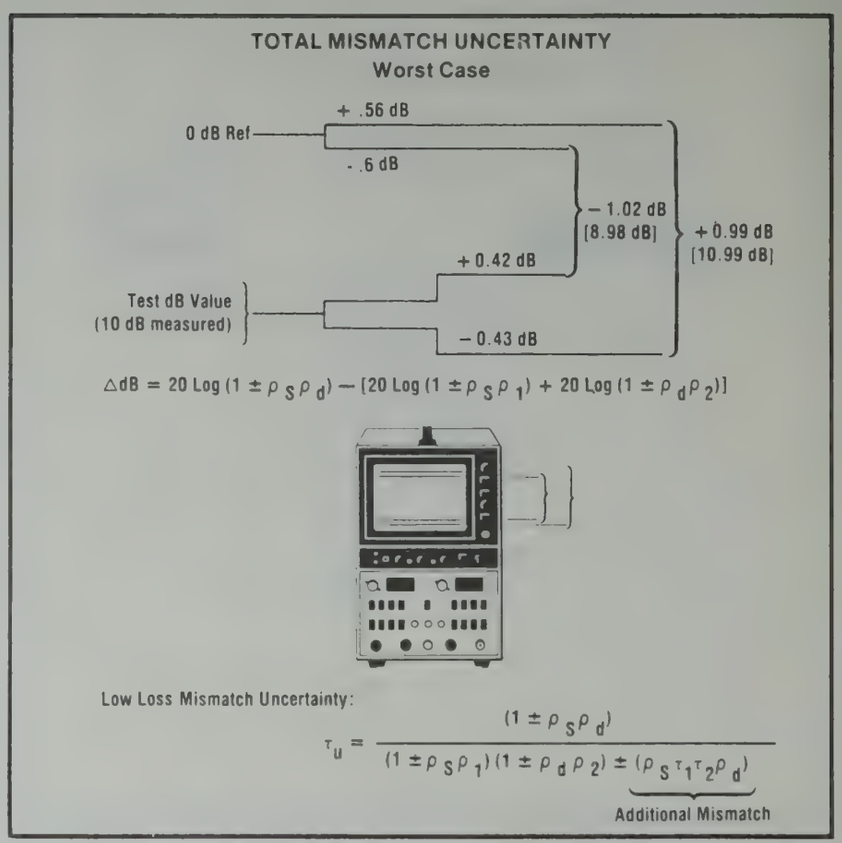
If the DUT has an input to output isolation of <6 dB, multiple reflections begin contributing to the uncertainty, and the error limits calculator cannot be used.

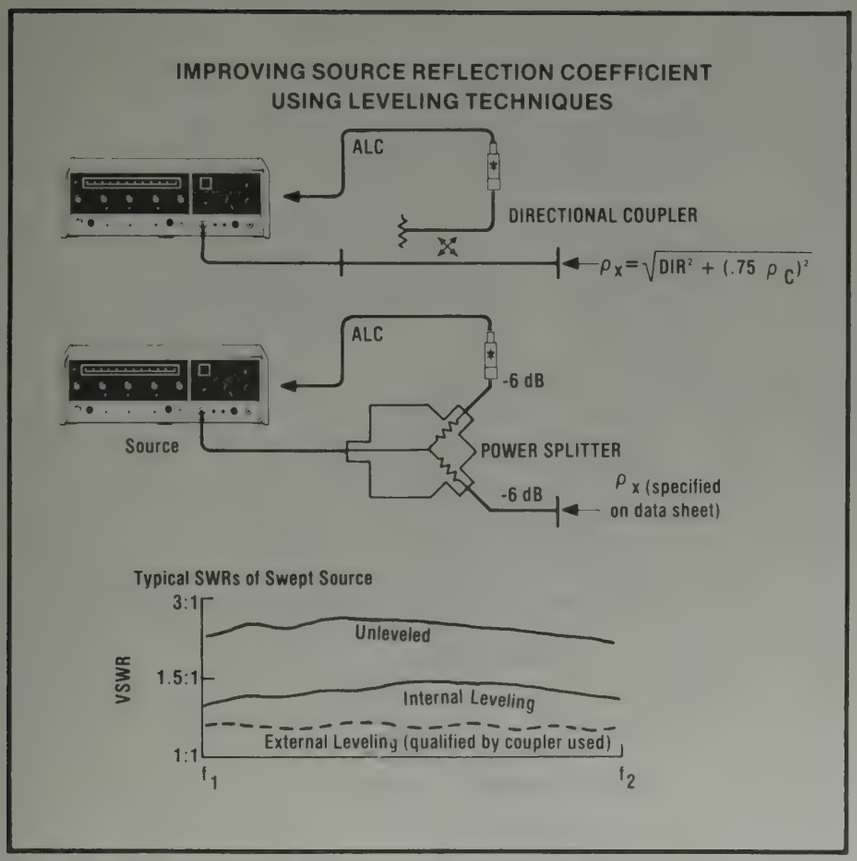


As the diagram shows, when the calibration window and the measurement window are combined, the total worst case uncertainty for a particular measuring system and DUT can be determined. Note that the positive measurement uncertainty is subtracted from the negative cal uncertainty and the negative measurement uncertainty is subtracted from the positive cal uncertainty. The reason for this can been seen mathematically from the original linear uncertainty equation; the measurement uncertainty is in the denominator, meaning it is subtracted when converted to dB.

Note also that the "Additional Mismatch" ($\rho_S\tau_1\tau_2\rho_d$) in the denominator of the linear equation is not present in the ΔdB equation. This can be done if the input to output isolation of the DUT is >6 dB. If not, the linear equation must be used to evaluate the uncertainty.

In the example, we see that the worst case uncertainty in measuring a 10 dB attenuator is approximately ± 1 dB! Obviously, we need to look at ways to improve this accuracy.

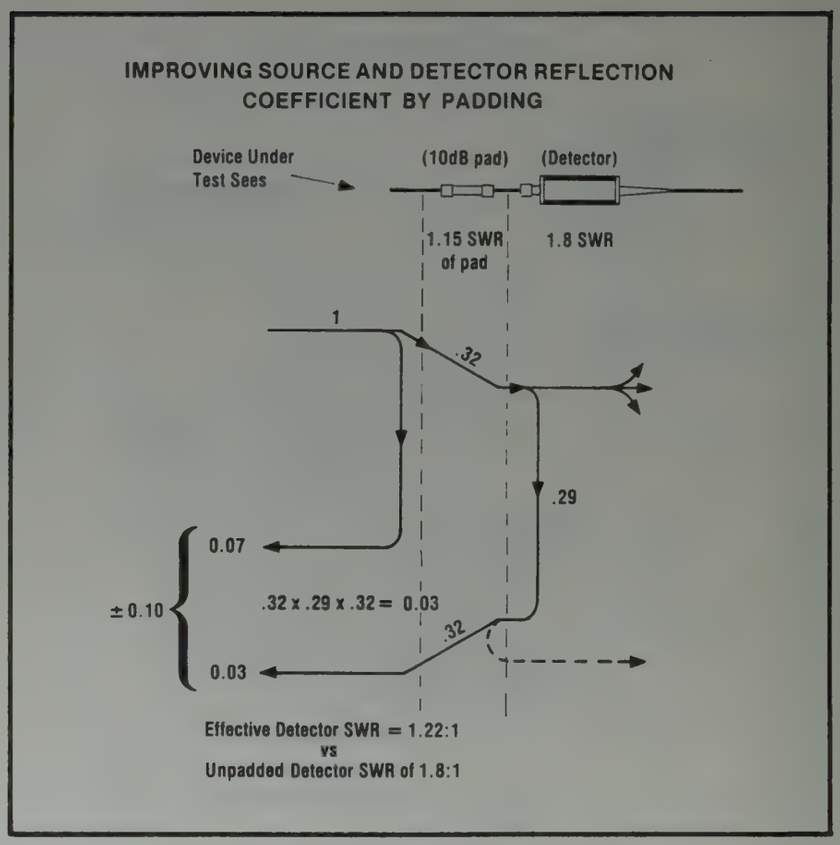

TOTAL MISMATCH UNCERTAINTY
Worst Case
0 dB Ref
+ .56 dB
- .6 dB
− 1.02 dB [8.98 dB]
+ 0.99 dB [10.99 dB]
+ 0.42 dB
Test dB Value (10 dB measured)
− 0.43 dB
$\Delta dB = 20 \text{ Log } (1 \pm \rho_S \rho_d) - [20 \text{ Log } (1 \pm \rho_S \rho_1) + 20 \text{ Log } (1 \pm \rho_d \rho_2)]$
Low Loss Mismatch Uncertainty:
$\tau_u = \frac{(1 \pm \rho_S \rho_d)}{(1 \pm \rho_S \rho_1)(1 \pm \rho_d \rho_2) \pm (\rho_S \tau_1 \tau_2 \rho_d)}$
Additional Mismatch



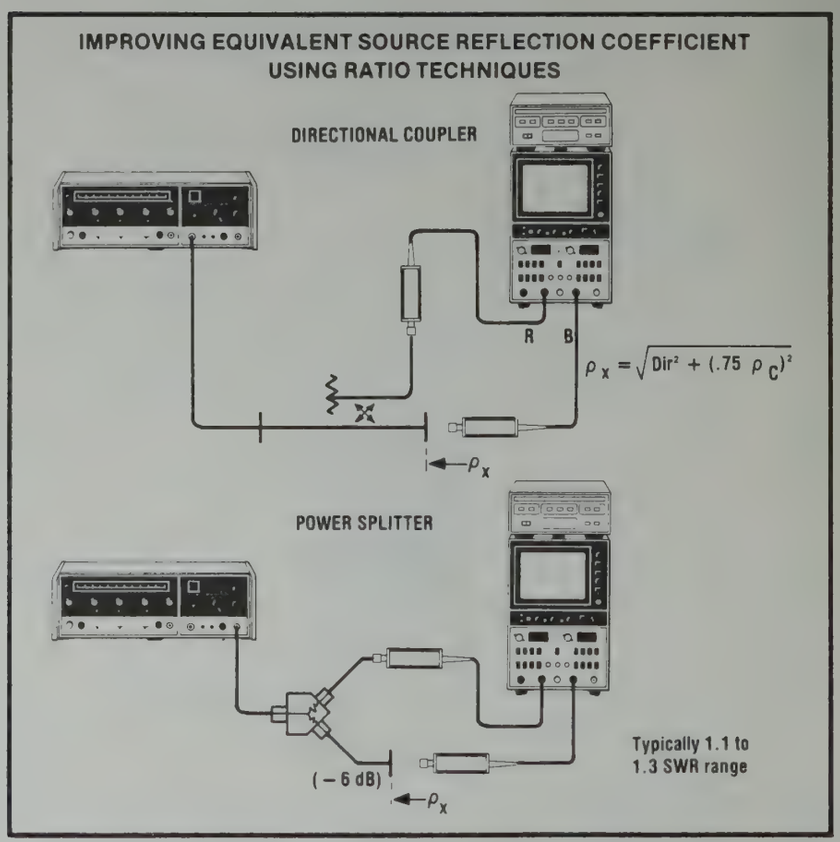
As discussed in the impedance section, leveling the source improves source match. In reflection measurements a leveling coupler when used with a similar reverse coupler improves frequency response; a dual directional coupler minimizes the number of connectors and their mismatches in addition. In transmission measurements the coupler may also be used; using a power splitter would improve the source match by a similar amount. The advantages of splitters include being very broadband because they are resistive and having extremely close frequency response tracking between the output ports ($< \pm .25$ dB from DC-18 GHz w/11667A). A potential liability is the insertion loss (6 dB) which may decrease the dynamic range.

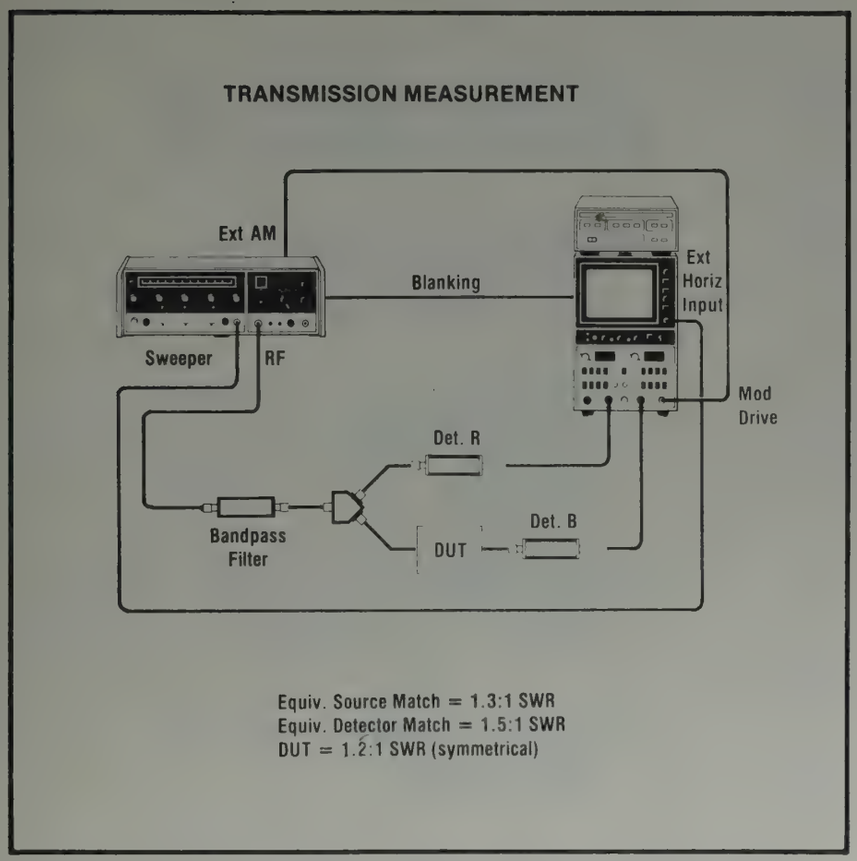
The effective source match using a coupler is the same as in the reflectometer set-up ($\rho_x = \sqrt{dir^2 + (.75\ \rho_c)^2}$. Using a Two-Resistor power splitter results in a source match as specified on the data sheet (1.1-1.3 for 11667A).

Another way of reducing mismatch uncertainty is by using pads (attenuators). A 10 dB pad can be used at the input to the detector (or at the output of an unleveled/unratioed source) to improve the match to essentially that of the pad. As the diagram shows, a signal traveling toward the detector first encounters the good match of the pad (e.g. ρ = .07). The signal is then attenuated by 10 dB before seeing the detector mismatch (.28 in this case). A signal level of .093 (.32 x .29) is subsequently reflected from the detector and experiences another 10 dB attenuation resulting in a signal level of .03 adding in some phase with the .07 reflected directly from the attenuator. So worst case this improves the detector SWR to 1.22 from 1.8. The major drawback to using this technique for improving source mismatch is that dynamic range may suffer.

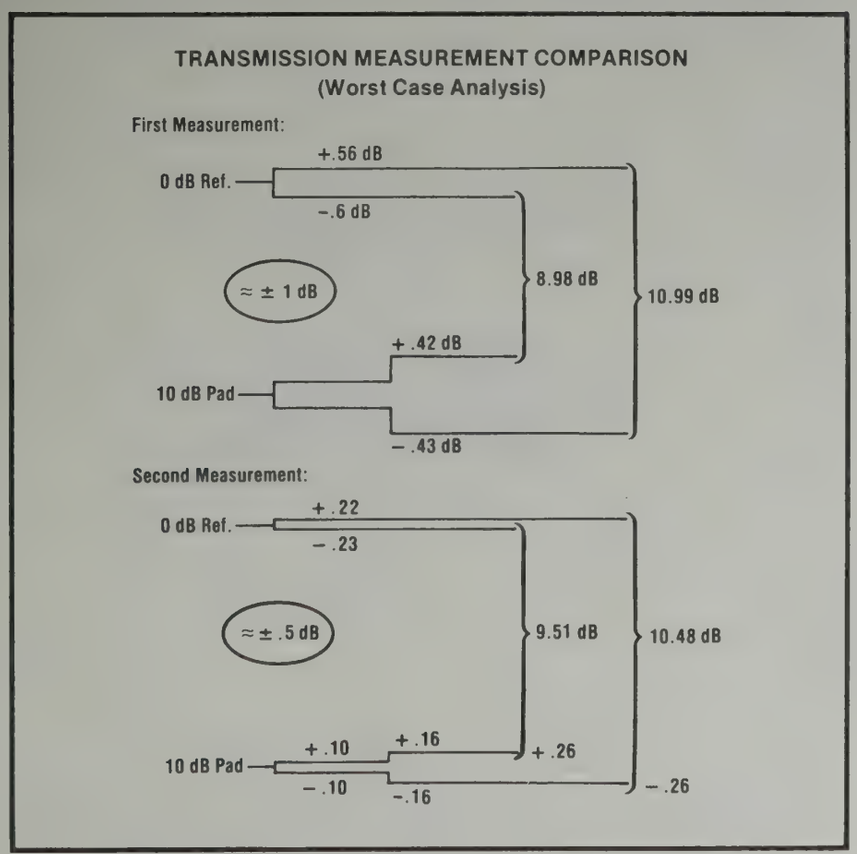


Just as in the reflectometer, a ratio measurement accomplishes the same improvements as does leveling. Again, the only caution is when measuring power sensitive DUT's.





This then is an improved transmission measuring system. Ratioing is used to improve source match to 1.3. In this case the detector was left unpadded because the dynamic range is required for other measurements. Note that a bandpass (or low pass) filter is used to minimize harmonics. This will be discussed later.



How much was the measurement improved by ratioing? The top brackets show the accuracy we obtained with the original simple measurement: approximately ± 1 dB. Using the Mismatch Error Limits calculator to calculate the improved uncertainty we see it is approximately ± .5 dB. Again, this is Worst Case mismatch uncertainty.

Note that the uncertainty is independent of the measured value of transmission as long as it is >6 dB (i.e., a 60 dB DUT would have the same uncertainty in dB as the 10 dB DUT). For higher attenuation measurements where more dynamic range is needed, removing the pad from the detector doesn't hurt the relative uncertainty much (including the pad would have decreased uncertainty to ± .3 dB).

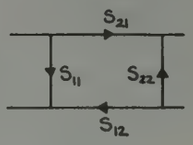
Why do we measure both magnitude and phase? Several important reasons will be discussed in this seminar. Magnitude and phase data are required for determination of complex impedance (both resistive and reactive components). All high frequency computer design programs require transistor S-Parameter data bands which of course means measurement of Magnitude and Phase for complete characterizations. Careful attention to both the magnitude and phase response of communications systems is essential for distortion free transmission.

Once the magnitude and phase of RF errors such as directivity have been characterized they can be removed from a measurement resulting in dramatic increases in measurement accuracy.

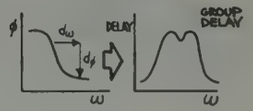
THE NEED FOR BOTH PHASE AND MAGNITUDE MEASUREMENTS

1. COMPLEX IMPEDANCE

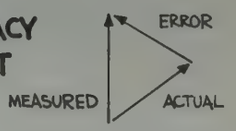
2. COMPLETE CHARACTERIZATION OF LINEAR NETWORKS



3. TRANSMISSION DISTORTION



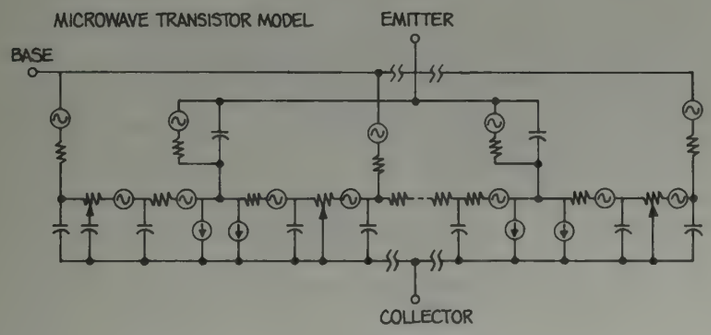
4. VECTOR ACCURACY ENHANCEMENT



Many of you may be familiar with S-parameters. For the benefit of those who aren't, we will have a quick review with emphasis on flow graphs. This is particularly useful when we talk about improving measurement accuracy. S-parameters are a tool used to characterize how a device modifies signal flow.

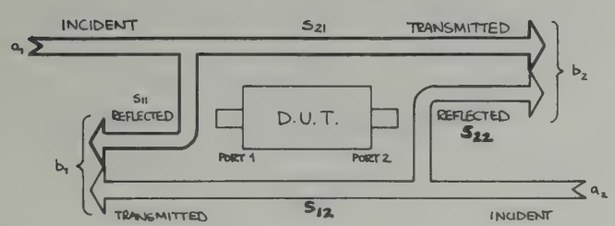
SCATTERING PARAMETERS

MODELING HIGH FREQUENCY NETWORKS



Who uses S-parameters? At low frequencies discrete element transistor models are very simple. However, as you can see, modeling becomes much more complex for high frequency devices. Circuit designers have addressed this problem by treating the transistor as a black box and then characterizing it at each frequency of interest. Today, this characterization process is most commonly accomplished by measuring the device's S-parameters. Once a transistor (or any other difficult-to-model-component) data file has been assembled, computer-aided-design techniques are commonly used to analyze the circuit. Using this exact modeling technique, microwave circuit designers have achieved phenomenal correlations between theoretical and actual designs.

SCATTERING PARAMETER DEFINITION



$$b_1 = S_{11} a_1 + S_{12} a_2$$

$$b_2 = S_{21} a_1 + S_{22} a_2$$

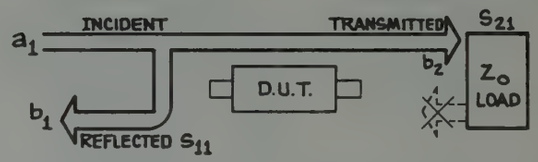
Let's look at a two-port device - that means it has two connectors - an input and an output. As a convention let "a" waves represent the energy wave entering a device at a particular port and "b" waves represent the energy leaving the device. At the input port, the "b" wave is composed of a portion of the incident wave that is reflected by the device, plus a portion of the incident wave on port 2 that is transmitted through the device. Likewise on the output port, the "b" wave at port 2 is equal to a portion of the incident wave at port 1 that is transmitted through the device plus a portion of the wave that is incident at port 2 that is reflected back out. Thus we have defined two simple linear algebra equations.

If we were to substitute current (i) for the "b's" and voltage for the "a's", we would have defined the more familiar Y-parameters. Although Y-parameters are analogous to S-parameters, they have few applications at microwave frequencies because of difficulty in obtaining the required open and short terminations.

How are the four S-parameters determined? S_{11} is the ratio b/a when a_2 is zero. How do we make a_2 zero? When the test device is terminated in a Zo load, no energy is reflected back into port 2 so a_2 is then zero. What is the physical meaning of S_{11}? It's magnitude value is identical to ρ (rho) or the devices input reflection coefficient. When combined with it's phase value, S_{11} is sometimes called the complex reflection coefficient (Γ). In a like manner the forward transmission coefficient S_{21} is determined.

S-parameters numbering convention is: The first number represents the port where energy is emerging from the device. The second number is the port where energy enters. So S_{21} means the ratio of the energy emerging at port 2 to the energy incident at port 1 (gain or loss is indicated by values greater or less than unity).

SCATTERING PARAMETER MEASUREMENT



$$b_1 = S_{11}\,a_1 + S_{12}\,a_2$$

$$b_2 = S_{21}\,a_1 + S_{22}\,a_2$$

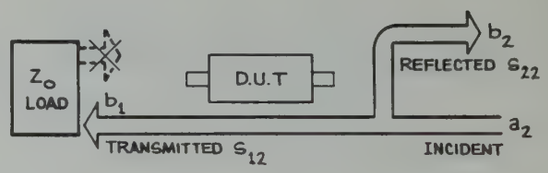
$$S_{11} = \frac{\text{REFLECTED}}{\text{INCIDENT}} = \frac{b_1}{a_1}\,\Big|\; a_2 = 0$$

$$S_{21} = \frac{\text{TRANSMITTED}}{\text{INCIDENT}} = \frac{b_2}{a_1}\,\Big|\; a_2 = 0$$

By placing the source at port-2 and terminating the input with a Zo load the a_1 term now becomes zero so that S_{22} and S_{12} can be determined.

Notice that the S-parameters are referenced to the Zo impedance of our network analyzer or most commonly 50 ohms. HP A/N 95-1 and A/N 154 give additional details on designing with and converting S-parameters to other useful quantities.

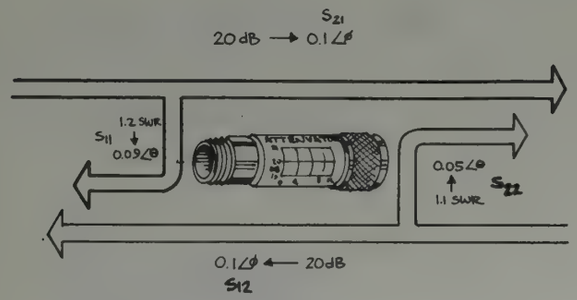
REVERSE MEASUREMENT



$$S_{22} = \frac{b_2}{a_2}\,\Big|\; a_1 = 0$$

$$S_{12} = \frac{b_1}{a_2}\,\Big|\; a_1 = 0$$

ATTENUATOR S-PARAMETERS

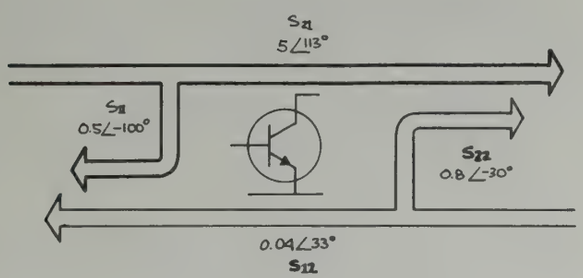


FREQUENCY DC TO 18 GHz

One of the nice things about S-parameters is that they are intuitive. That is not true for Y-parameters at high frequencies. As an example: Y_{21} is the forward transadmittance or the ratio of the current at port 2 over the voltage at port 1 with port 2 shorted. It's very difficult to relate to that parameter whereas S_{21} is just the gain or loss of the device expressed in linear form. With this 20 dB attenuator, we see that our familiar measurement of loss on the attenuator transforms to a S-parameter voltage ratio of 0.1. Since these are complex parameters, there is an angle associated with each S. Likewise the more familiar measurement of SWR on the device translates over to an S_{11} or input voltage reflection coefficient of .09. Basically we have been making S-parameter measurements all along. We have just used different names for the same quantities.

This attenuator is a bi-lateral (reciprocal) device, i.e. it has approximately the same loss whether you drive it from port 1 or from port 2. ($S_{21} \approx S_{12}$)

TRANSISTOR S-PARAMETER



HP HXTR 6101 @ 1 GHz

A microwave transistor is a non-bilateral device; its reverse transmission coefficient is markedly different from it's forward coefficient. Also, its S-parameters vary rapidly with frequency so the S_{21} of 5 (14 dB) applies only at 1 GHz. Also, its reflection coefficients are very high, as is typical of small signal high frequency transistors.

So why do we find S-parameters used so predominantly at microwave frequencies?

1. They relate to familiar things such as gain, loss, and reflection coefficient.

2. When used in circuit analysis programs they describe exactly how a characterized component will perform.

3. They are easily measured with conventional reflection and transmission measurement systems.

4. They lend themself to analysis as we shall now see as we now review flow graph techniques.

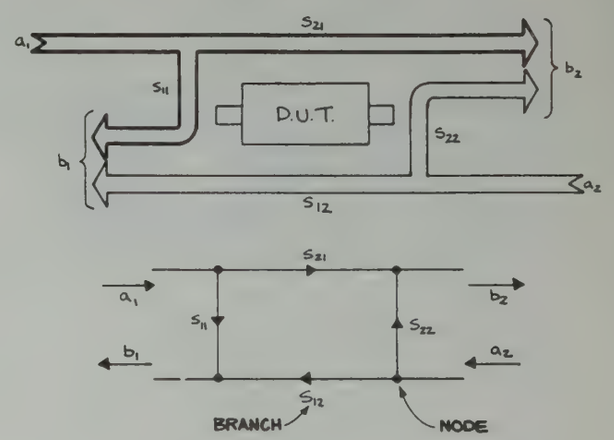
WHY S-PARAMETER?

1. CONCEPTUALLY SIMPLE
2. EXACT DATA
3. EASY TO MEASURE
4. ANALYTICALLY CONVENIENT
   -FLOW GRAPH ANALYSIS-

This is a flow graph representation of a two-port device. The input port has two nodes: one representing the entering or "a" wave and one for the emerging "b" wave. Lines that connect nodes are called branches. Each branch has an arrow and value corresponding to a S-parameter. Energy will only flow in the direction of the arrow.

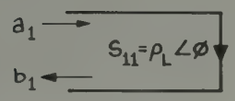
How many S-parameters does a device have? Square the number of ports (connectors) and you have the number of S-parameters. Thus, a two-port has four S-parameters.

NETWORK FLOW GRAPHS
TWO PORTS

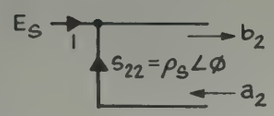


EXAMPLE FLOW GRAPHS
ONE PORT

LOAD



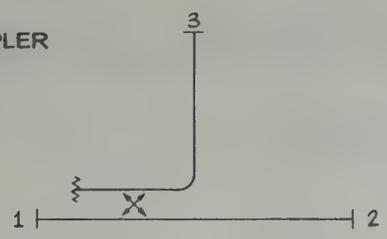
SOURCE

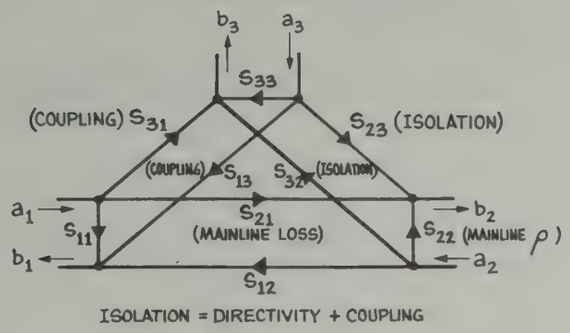


Now a one-port device has one squared or one S-parameter. Here are several one-port examples: a load and a source. Each flow graph shows how the incident "a" wave will be modified and re-emerge as a "b" wave.

EXAMPLE FLOW GRAPH
THREE PORTS

COUPLER



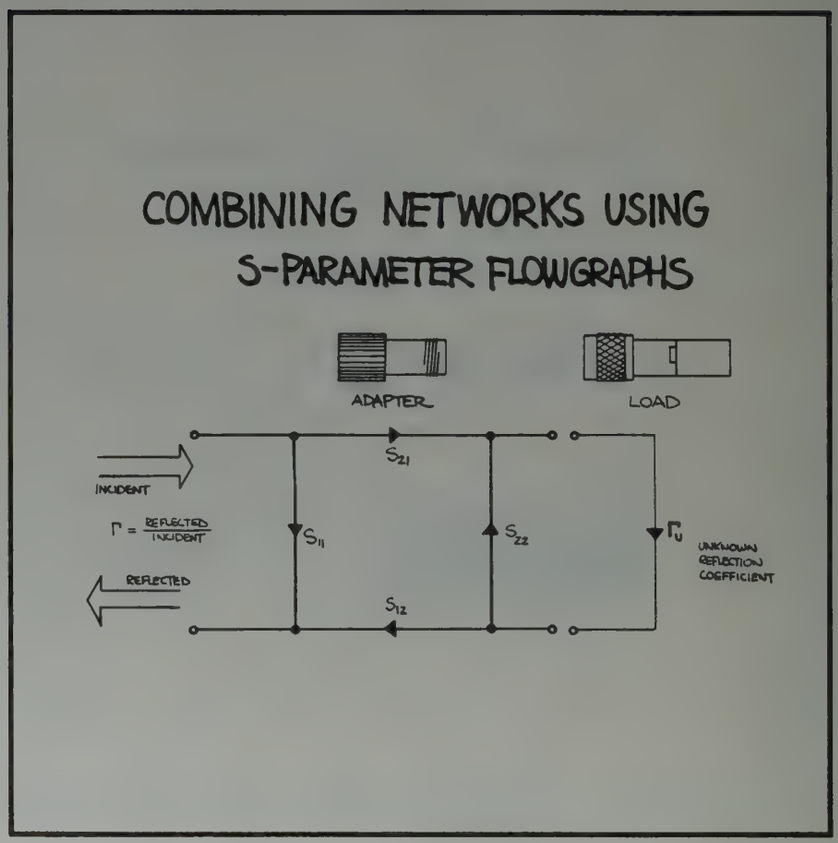


ISOLATION = DIRECTIVITY + COUPLING

A three-port has 3 squared or 9 S-parameters as shown by this flow graph representation of a coupler.

The portion of energy incident on port one that is available at port three is S_{31} or the coupler's coupling factor. The portion of energy incident on port two available at port three should be zero for a perfect coupler. Of course it is very difficult to build a coupler whose S_{32} or directivity is zero. S_{22} is the mainline reflection coefficient.

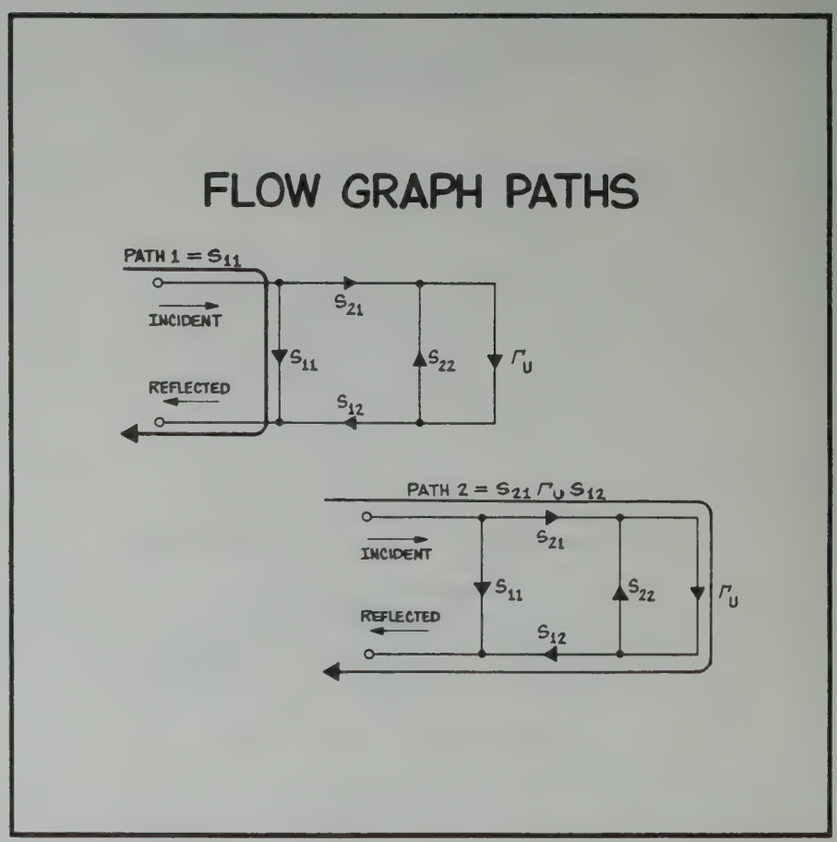
Flow graph techniques can be used to determine the overall response of a network once the component parts are known. As an example, let's use an adapter to convert a Zo termination to another connector type. We know intuitively that the load isn't improved by putting something in front of it. We will show how to predict exactly the "new" load reflection coefficient once we have measured the adapter and load S-parameters.



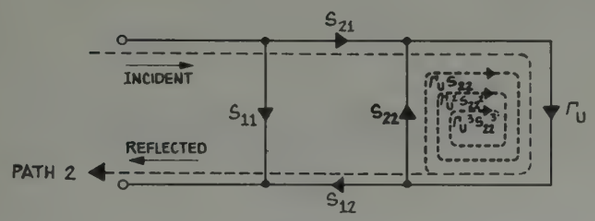
A flow graph reduction technique known as "Mason's Nontouching-loop rule" (see bibliography) say that the overall response of this network is the sum of paths which energy can flow through the network.

Energy flows from source to sink in the direction of the arrowhead. Thus one obvious path for energy incident at port one is the direct path through the adapter's S_{11}.

A second path or direction energy may flow is through S_{21}, then through the gamma of the load, and then back out through S_{12} of the adapter. The combined value of this path is the product of each branch or ($S_{21}\Gamma uS_{12}$). Notice that while energy can't flow out of S_{21} into S_{22}, nothing prevents it from coming up S_{22} and back into Γu.



FLOW GRAPH LOOP



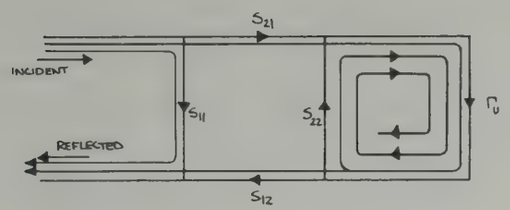
$$\text{LOOP} = 1 + \Gamma_U S_{22} + (\Gamma_U S_{22})^2 + (\Gamma_U S_{22})^3 + \ldots$$

$$= \frac{1}{1 - \Gamma_U S_{22}}$$

$$\text{PATH 2 (MODIFIED BY LOOP)} = \frac{S_{21}\, S_{12}\, \Gamma_U}{1 - \Gamma_U\, S_{22}}$$

Some of the energy does flow in that direction and recirculates in a loop. Each time it transverses the loop a product of the coefficients is generated, or an infinite series. A mathematical identity converts the infinite series to a simple expression which is multiplied by the S_{12}, S_{21}, and Γu branches to give an overall value for path two.

COMPOSITE RESPONSE



$$\frac{\text{REFLECTED}}{\text{INCIDENT}} = \text{PATH 1} + \text{PATH 2 (MODIFIED BY LOOP)}$$

$$= S_{11} + \frac{S_{21}\, S_{12}\, \Gamma_U}{1 - \Gamma_U\, S_{22}}$$

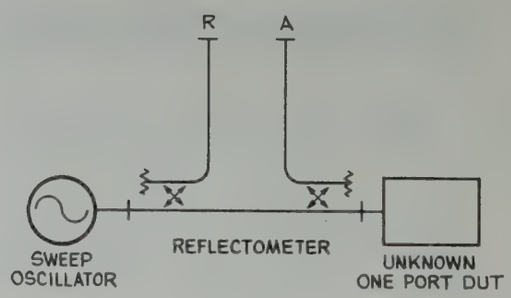
Now for the sum. This simple mathematical relationship is an exact solution to any two-port network terminated in a one-port. In our example, it is the reflection coefficient of the adapter/load combination. Notice the prominance of the S_{11} terms. No mater how low the load reflection coefficient is, the adapter's input reflection coefficient (S_{11}) will seriously degrade the "new" load even when it is comparable in value to gamma "L".

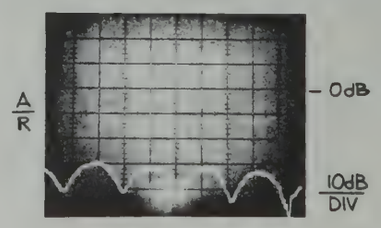
We've reviewed S-parameters and flow graphs as a basis for understanding vector accuracy enhancement techniques. Particularly in high frequency network analyzer measurement these techniques can have a very significant effect on measurement accuracy.

ACCURACY ENHANCEMENT IN VECTOR MEASUREMENT SYSTEMS

Here is a typical display for a reflection measurement of a one-port test device. The test device is a near Zo termination. What causes the peaks and valleys in the results? Not the device under test! A thin film resistor just couldn't have these rapid variations with frequency. Unfortunately these very large variations (greater than 20 dB in return loss) are caused by imperfections in our measurement systems.

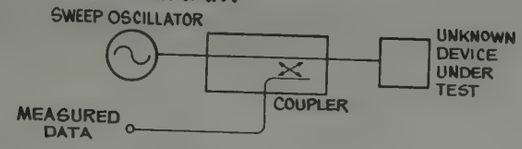
THE PROBLEM

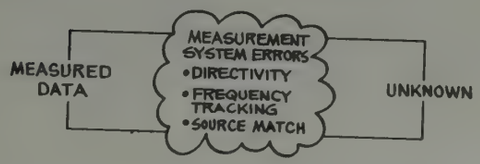




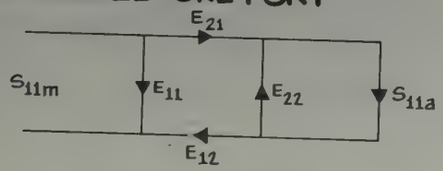
REFLECTION ERROR MODELING

BLOCK DIAGRAM





ERROR MODEL - ONE PORT



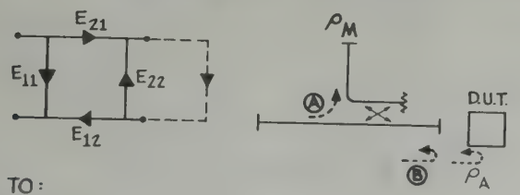
$$S_{11m} = E_{11} + \frac{E_{21} E_{12} S_{11a}}{1 - E_{22} S_{11a}}$$

Measurement system errors prevent measured data from being a true representation of an unknown. By understanding where these errors come from and then by developing an appropriate model, we can remove their effects from a measurement. We will show that a reflection measurement can be modeled with this same flow graph we previously developed. It should not be a surprise that the mathematical solution is in exactly the same form. Only the variable names have changed.

Recognize however that in any modeling of a measurement system there are some errors that cannot be modeled. For example, drift and noise are not capable of being represented in a systematic way. Likewise, residual FM, the repeatability of the individual connector type, or test cable variations as we move to connect it to our device are not modelable since they are not repeatable. We'll have a look in detail at what each of these E-parameters or error coefficients, in our model represent.

DIRECTIVITY ERROR

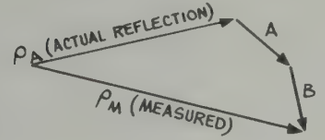
LEAKAGE SIGNAL INDEPENDENT OF DEVICE UNDER TEST



DUE TO:

A - IMPERFECT COUPLERS OR BRIDGES

B - REFLECTION FROM COUPLER CONNECTORS, ADAPTERS OR TEST CABLES



First examine E_{11}. This branch of our model represents the signal that never gets to the device under test. It leaks off or reflects before it gets there due to:

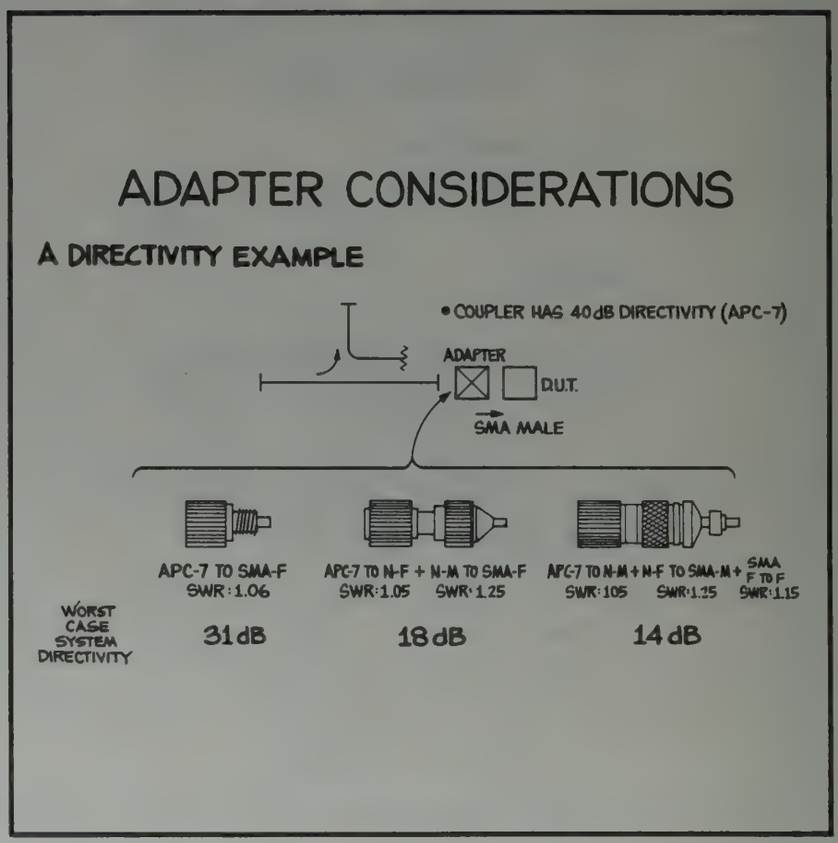
1. The imperfection of the signal separating device.

2. The adapter or test cable that physically connects to the device under test.

We see these errors combine vectorally with the true reflection coefficient (ρA) to give an inaccurate value for the measured date. We call the combined vector errors: Effective Directivity.

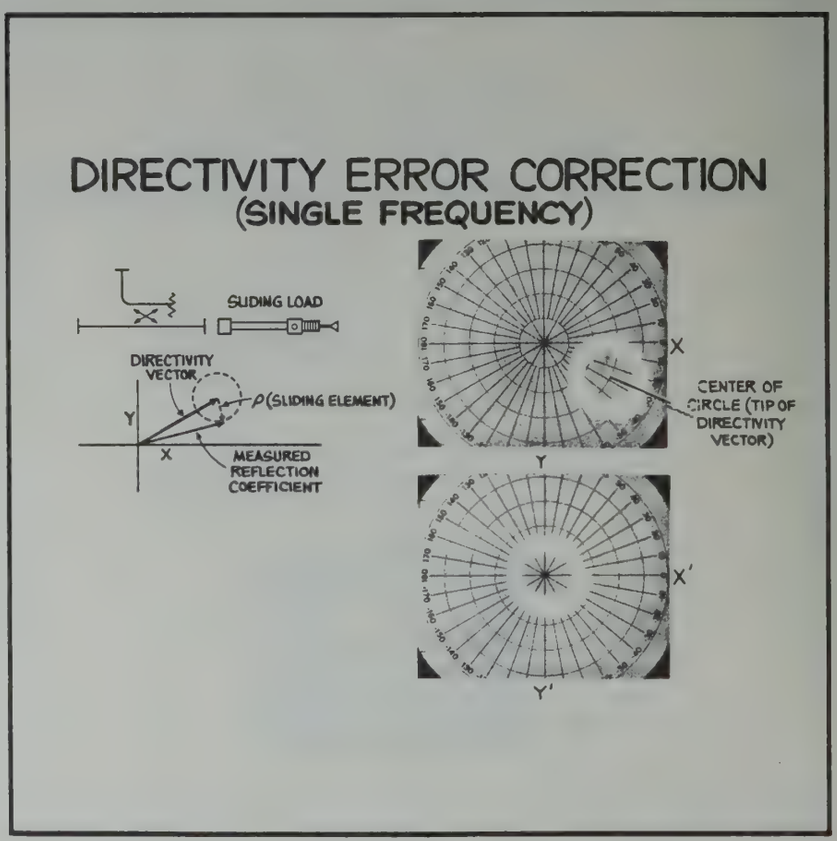
Let's consider the effect of adapters. In this case, we're trying to measure a device that has an SMA male connector on it. Our coupler is very good. At 18 GHz it has a 40 dB directivity in APC-7. We need to adapt from APC-7 to SMA. Obviously, any adapter will degrade our measurement. Here are three different approaches to adapting.

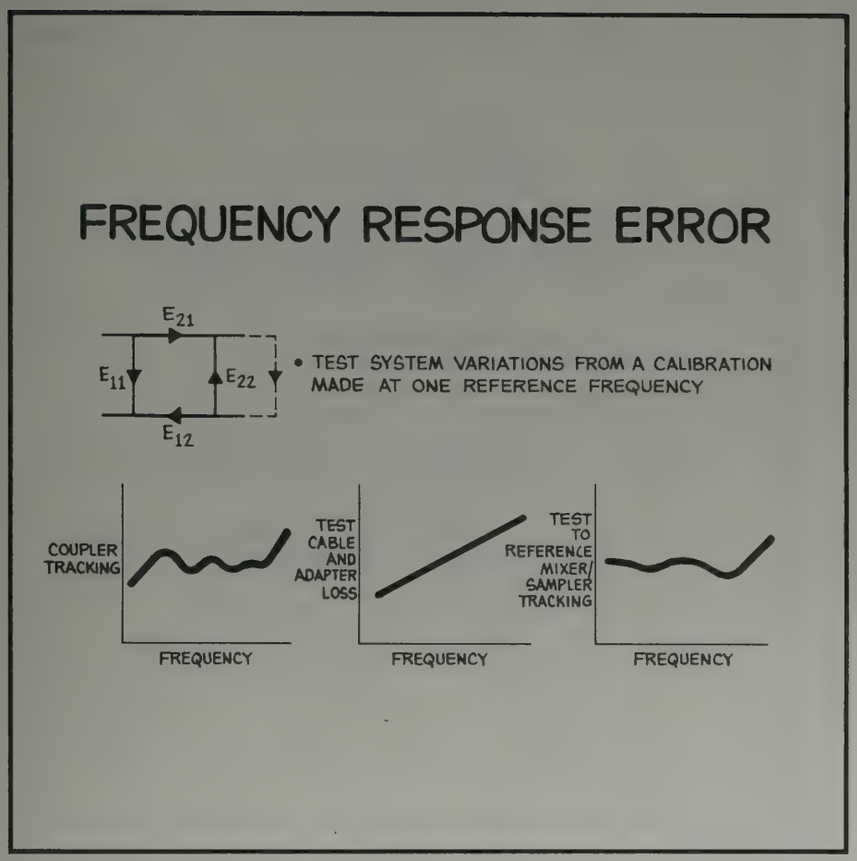
The best way, of course, is to buy the finest APC-7 to SMA female adapter. However, even that, at 18 GHz, reduces the system directivity to 31 dB. But, sadly enough it's common to see combinations of two and even three adapters. The resultant performance of these combinations can be as bad as 14 dB effective directivity. Remember you can't measure a return loss any better than the system's directivity.



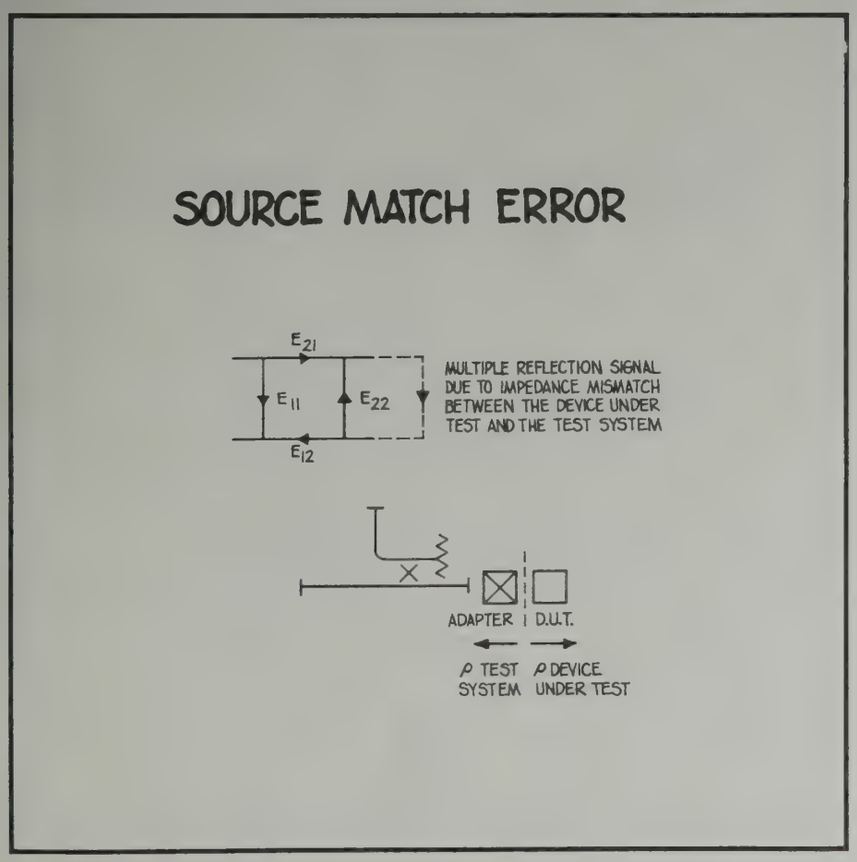
How can we remove some of these directivity errors? There has been a method available for many years with manual vector measuring systems. A sliding load can calibrate the total measuring system including adapters at a single CW frequency. As we physically slide the load, we describe a circle on the screen. The radius of that circle is the reflection coefficient of the actual sliding element within the sliding load. The mechanical ratio of the inner to outer diameters, the precision nature of the sliding load, represents the center of that circle. In APC-7 or Type N this is normally equivalent to a return loss of greater than 52 dB. Note that on a system the center of the circle is not the center of the polar display. The directivity error is the offset or vector from the center of that circle to the center of the display. At a single frequency. We take the X and Y positions and offset until we set it to the center of the polar display. In effect, we've stored the vector directivity error at the one frequency in the settings of X and Y position.

This can be a laborious process when done manually at several frequencies, but is easily accomplished by automatic systems.





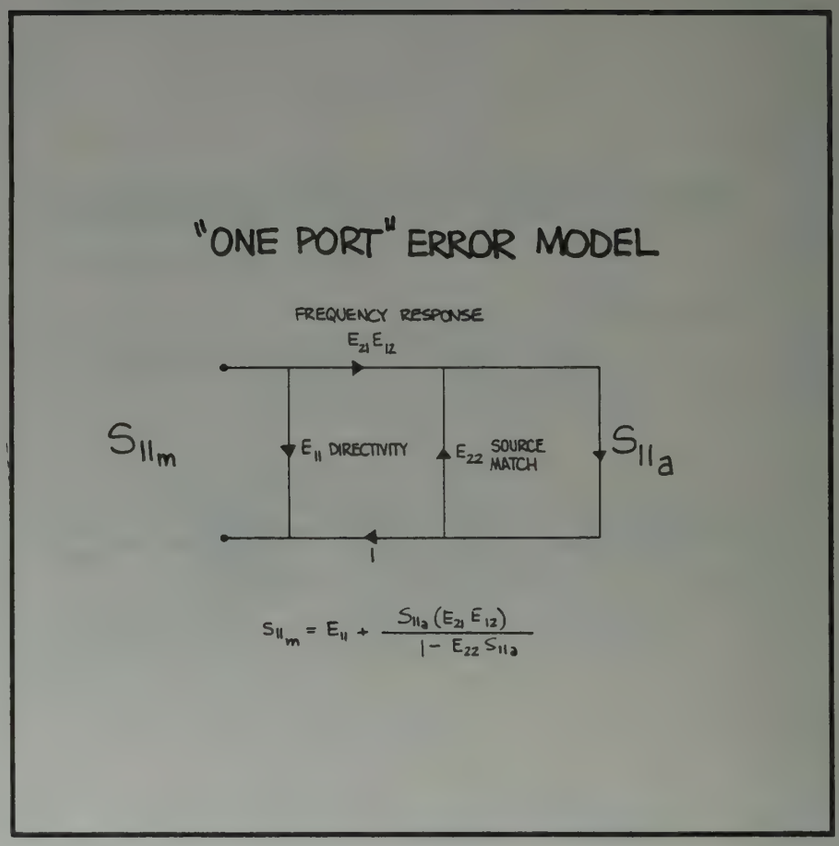
Let's go back to our model and take a look at some of the other error components. The E_{21} and E_{12} represent the loss in our system going to and coming from the DUT. This error called frequency response is actually a composite of many of the components of the system: the couplers, test cables, and mixers.



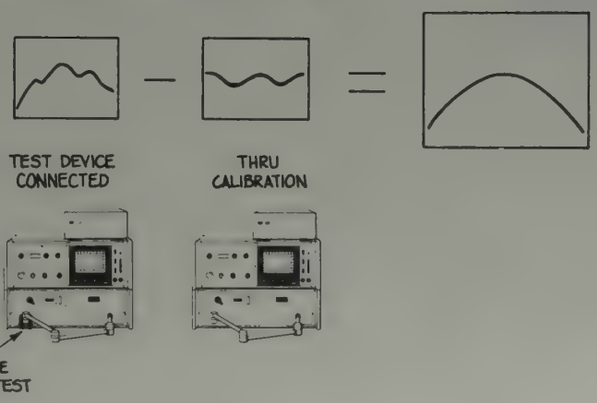
The last error coefficient (E_{22}) to consider in our analysis of a reflection measuring system is something called source match. When we look back into a measuring system including all it's adapters and cables, we do not see an ideal match. When a measuring signal reflecting back off the device under test comes back to the test system, an error loop is formed with signals bouncing back and forth. The error is related to the product of these coefficients so it's really only a major effect when the device we're testing has high mismatch. In the case of semi-conductors or a filter measured out of band it can be a major error approaching 10%.

Let's go back to our complete model again and its now familiar equation. Notice the E_{21} and E_{12} terms appear together as a product since we can really consider them as one factor called frequency response. E_{11} is the equivalent directivity of the system and E_{22} is the source match. They are the three error terms whose effects we want to remove. We need a mechanism whereby we can determine a set of coefficients for a particular measurement setup and store those vector coefficients so that later we can remove them from our measured data. Since there are 3 unknowns in our equations, we require 3 devices whose characteristics are known to solve this equation.

Now we're going to "calibrate" using a set of devices we recommend. However, any 3 known devices will suffice theoretically.

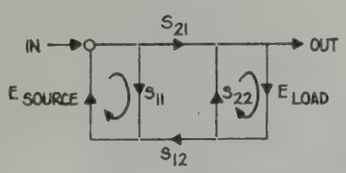


TRANSMISSION FREQUENCY RESPONSE CORRECTION



Before we consider two-port reflection further, let's consider transmission. Most of us are familiar with normalization. A through is connected and the frequency response is stored either using some digital form of storage or a grease pencil. Then we insert the device under test and subtract the frequency response data from the total measured transmission data. The resulting data is corrected for frequency response.

TRANSMISSION MISMATCH ERRORS



INTERACTION BETWEEN DEVICE UNDER TEST AND TEST SYSTEMS SOURCE AND LOAD IMPEDANCE

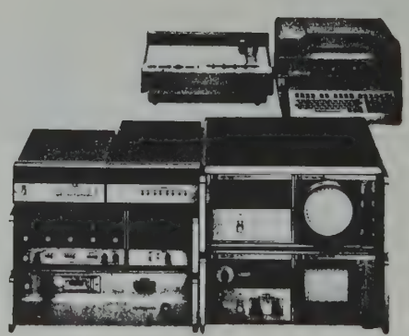
$$\text{FORWARD TRANSMISSION (MEASURED)} \approx \frac{S_{21}}{1 - E_S S_{11} - E_L S_{22}}$$

When the test system source and load impedances are not exactly equal to Zo (50 ohms) they will interact with the two-port we are measuring and cause a transmission mismatch error. This error effect is very similar to the one-port mismatch previously discussed.

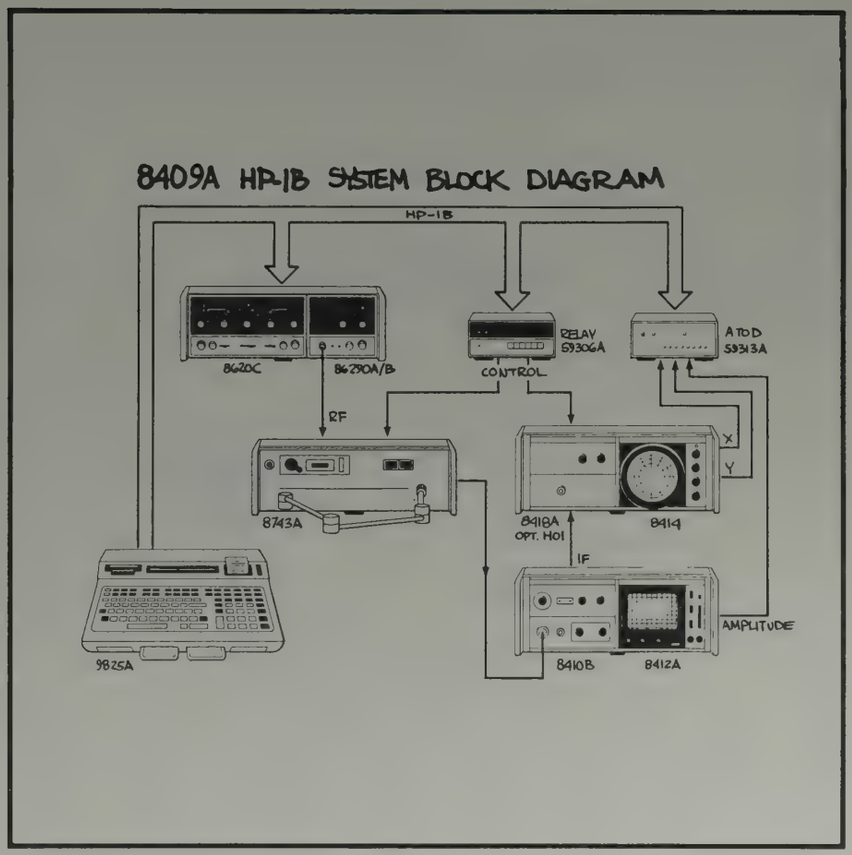
Let's take an example. Let's say the test system matches are quite good (1.2 SWR or .09 reflection coefficient) but we are measuring a highly reflective device like a filter at its 3 dB down frequency (a reflection coefficient of 0.7). It can be shown that these measurement conditions result in an error approaching 15% or >1 dB out of 3.

The measurement system shown here is controlled by a desk top computer. The controller automatically sets frequency, programs the state of the instruments, and reads data from the network analyzer. It also performs in a several hundred milliseconds, calculations which would take hours when done by hand.

AN AUTOMATIC NETWORK ANALYZER

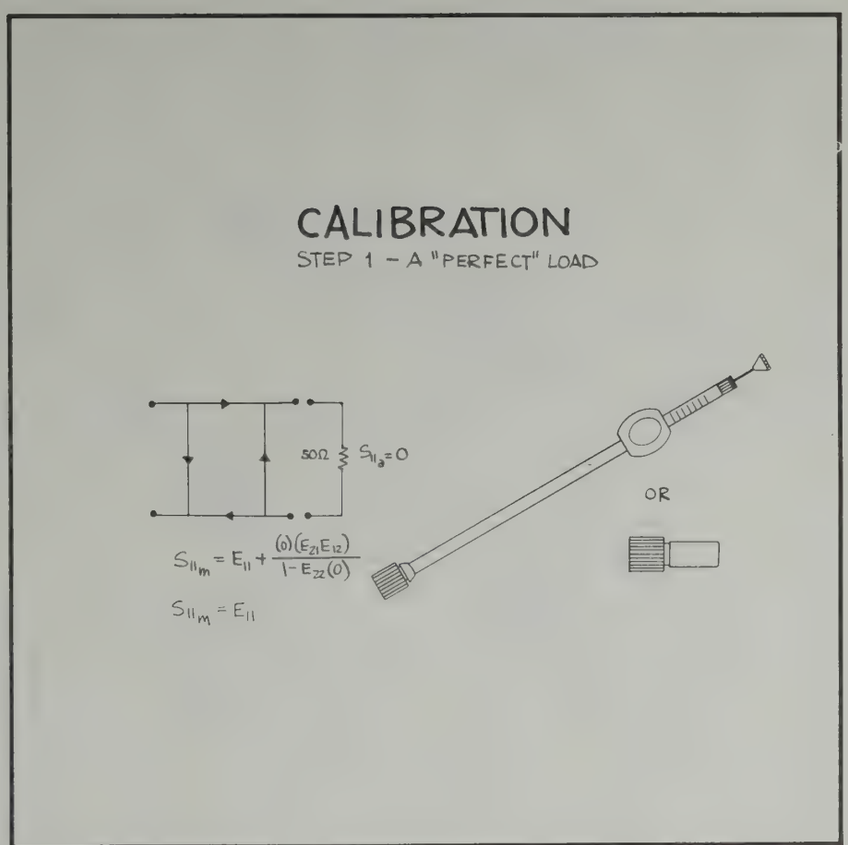


Looking at the block diagram of this measurement system, let's trace the path of the RF energy in our system. The signal feeds from the RF port of the test oscillator (2 - 18 GHz in this configuration) to the test set. There a sampler converts reflected or transmitted signals to an IF frequency which is applied to the 8410 Network Analyzer. For accuracy considerations two displays are used: log amplitude is obtained from the phase-magnitude display and phase from the polar display. This system is documented in HP Application Note 221.



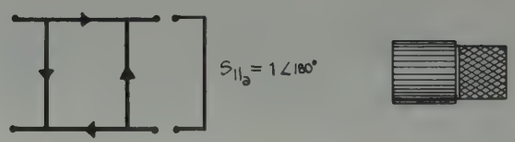
Suppose we start our calibration with a "perfect" load. Now this might be a sliding load or at lower frequencies, it might be a fixed termination in either 50 or 75 ohms or wave guide - whose reflection is known to be extremely low. If we substitute zero for the value of S_{11a}, the second term drops out and the measured data is the directivity.

The error is reduced to the uncertainty associated with the "standard" load. With a sliding load, several positions (4 - 6) of the termination are set at all frequencies. At the last setting the center of the data points is computed for each frequency. Thus the quality of the moving termination within the sliding load does not effect measurement accuracy. The accuracy is determined by the precision of the geometry of the air line section preceding the moving termination (typically >50 dB return loss).



CALIBRATION

STEP 2 – SHORT



$$S_{11m} = E_{11} + \frac{(-1)(E_{21}E_{12})}{1-E_{22}(-1)}$$

But we still have the source match and the frequency tracking coefficients to determine. We need two known devices. The first is a short circuit. This establishes our measurement reference plane. We temporarily store the results.

CALIBRATION

STEP 3 – OPEN

$$S_{11m} = E_{11} + \frac{(1)(E_{21}E_{12})}{1-E_{22}(1)}$$

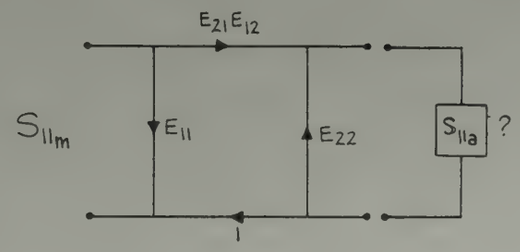
The final standard can be an open circuit (whose outer conductor is extended to prevent radiation). The coaxial open has a small fringing capacitance whose value is a function of the geometry of the connector type. This capacitance is simply and accurately modeled. For waveguide we can use an offset short circuit, i.e. a short preceded by a known length of line.

With this final independent solution we can now compute both source match and frequency tracking coefficients.

Our previous error equation can be re-written in this form. Notice that the actual unknown value is now a function of the three error coefficients and the measured value of the unknown (S_{11m}). This is the basis of vector accuracy enhancement. The three error coefficients are calculated for each frequency of interest and stored in a matrix (six numbers per frequency). As part of the measurement process, S_{11m} is measured and this equation calculated at each frequency of interest.

Now all of this could be done manually where we write down the data and solve each of these equations, but it's far easier when we use some form of automation.

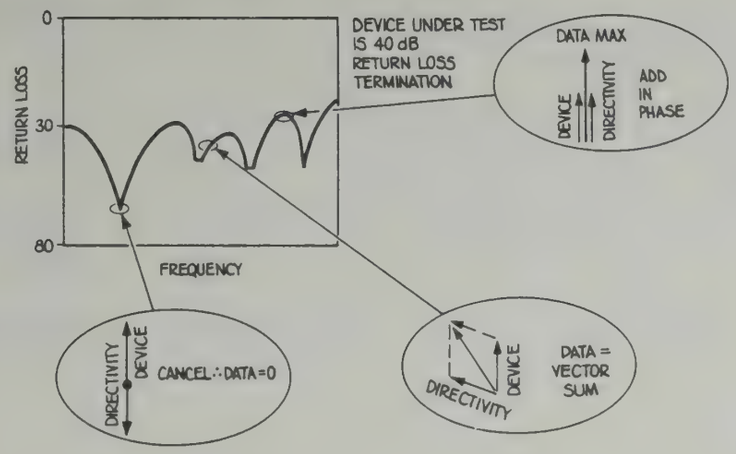
NOW CONNECT THE DEVICE UNDER TEST



$$S_{11a} = \frac{S_{11m} - E_{11}}{E_{22}(S_{11m} - E_{11}) + E_{21}E_{12}}$$

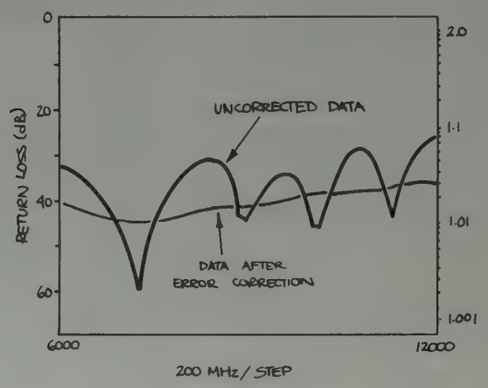
Let's take an example of what can be accomplished by this error correction technique. Again, we measure a good termination. Initially the system is programmed to output a plot of data that is uncorrected. We notice the familiar ripple effect that's caused by the interaction of the directivity error signal with the reflection signal of the device. At the peaks the devices reflection signal adds in phase with the directivity signal producing a data maximum. At the low valleys, the directivity signal and the device data are approximately the same magnitude but out of phase so they cancel each other and give a return loss approaching infinity. Directivity and device data combine somewhere between out-of-phase and in-phase between the valleys and peaks.

UNCORRECTED DATA



This is what the corrected data looks like. Notice that the data is much more realistic, and has an absolute (smoother) level of approximately 40 dB return loss. Our limit at the low reflection end of our measuring capability is set by the standard used for the "perfect" load.

BEFORE AND AFTER CORRECTION



In summary, our reflection error correction (one-port) technique has:

1) Extended the effective system directivity down to the return loss of sliding load.

2) Included the effects of adapters and cables.

3) Incorporated the source match uncertainties so important when measuring high reflection devices.

If we were measuring reflection on a two-port device we can use this same "one-port" method and terminate the two-port's output with a "perfect load". However, this can lead to substantial errors if the termination is imperfect and the two-port has low loss.

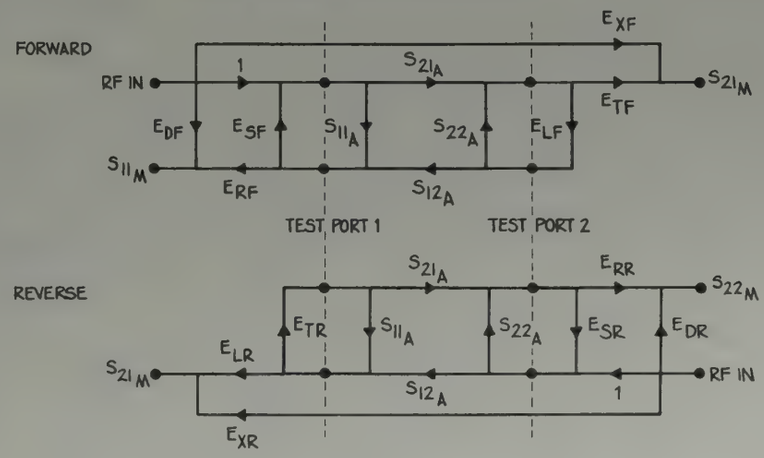
CONTRIBUTIONS OF REFLECTION ERROR CORRECTION

- EXTENDS SYSTEM DIRECTIVITY
- REMOVES EFFECTS OF ADAPTERS
- REMOVES SOURCE MATCH UNCERTAINTIES

Transmission mismatch errors can be removed with a more complete error model. We are considering here the use of a two-port test set which can reverse the flow of test signal power without having to physically turn the two-port around.

The reflection calibration terms are similar to the one-port except for the fact there are two sets of three coefficients for both forward and reverse. Transmission tracking, leakage and load match terms are added for both directions resulting in 12 terms for each frequency.

COMPLETE CALIBRATION MODEL
AN EXAMPLE (8409A)



Even after vector error correction, there are some residual errors that because of their nature or because of the complexity involved are not modeled. This is a typical error budget for the HP 8409A system where after RMS addition we have about a 0.1 dB error in a low loss measurement.

WHAT'S LEFT AFTER ERROR CORRECTION?
AN EXAMPLE (8409A)

	TYPICAL ERRORS		MAGNITUDE (dB)	PHASE (DEG)
• REPEATABILITIES:	TEST SET SWITCHES (8743)		.04	0.2
	CONNECTOR (SMA)		.03	0.2
	TEST CABLE		.02	1.0
• DETECTOR ERRORS:	MAGNITUDE (8412)	UP TO 40 dB	.02 dB/dB	
	PHASE (8414)	UP TO 40 dB		1.5
• HARMONIC SKIP:	ABOVE 1 GHz		0.1	0.5
	NOMINAL ERROR AT 1 dB		.115 dB	1.89° (RMS)
	AT 10 dB		.23 dB	

A network analyzer's basic magnitude and phase measuring capability can be used in place of dedicated one-of-a-kind test systems in many situations. Its application is often limited only by ones imagination in routing the input signals as we shall see from some of these actual customer examples.

Note that some of the more interesting of these network analyzer applications can provide non-linear parameter information with this "linear only" measuring tool. It should also be pointed out that most of these applications have been or could be automated using computer controllers even though that aspect is not shown in the following block diagrams.

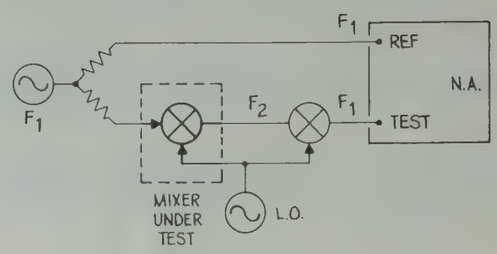
MEASUREMENT APPLICATIONS

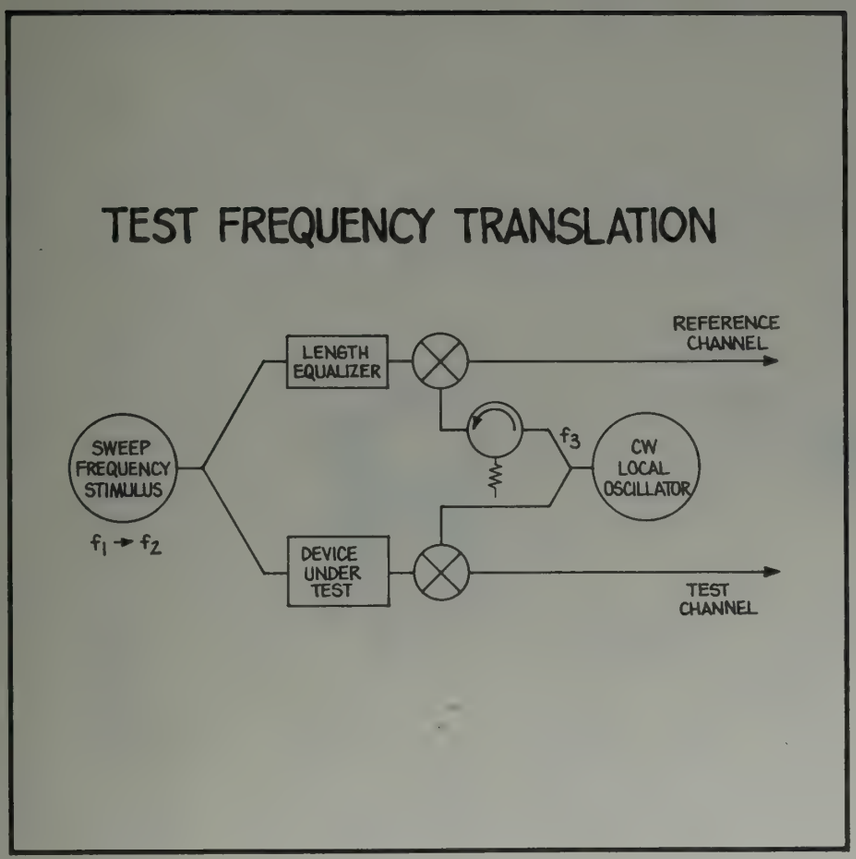
In mixer testing, it should be remembered that the network analyzer's reference and test channel signals must both be the same frequency. In spite of this restriction many users have successfully used network analyzers to characterize translators particularly when phase match is required. If the local oscillator drive is available we can drive a second translator to recreate the original frequency F_1. Often this second translator can be assumed distortionless or we are only interested in measurements relative to a "standard" mixer.

A network analyzer's narrowband receiver is essential to this class of measurements because it prevents the detection of harmonic and spurious responses.

FREQUENCY TRANSLATOR CHARACTERIZATION

WHEN LOCAL OSCILLATOR IS EXTERNAL





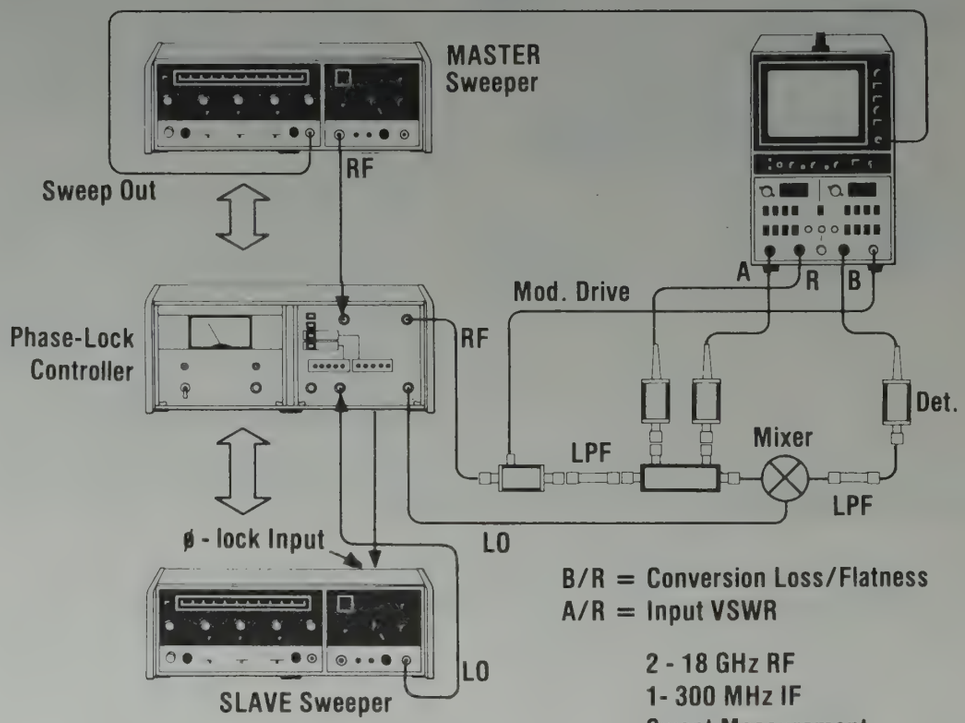
It is often possible to use a pair of mixers to translate the operating frequency range of a network analyzer for testing at higher frequencies.

The CW local oscillator frequency is normally set below the lowest frequency of the swept frequency stimulus. In this way the sign of the phase measurement is correct. If the local oscillator is tuned above the band being swept the sign of the measured phase angle must be reversed. For example, using the HP R8747A Transmission and Reflection Test unit we may down convert any 2 GHz window in the 26.5 to 40 GHz frequency range. To measure over the 38 - 40 GHz band set the L.O. to 36 GHz for correct phase data. The translated signal fed to both reference and test channels of the 8410 then varies from 2 to 4 GHz.

This class of systems usually trade offs some dynamic range for this higher frequency operating range due to the conversion loss of the mixers. A particular problem in attaining higher dynamic ranges is the cross coupling between the two mixers through their local oscillator ports. The isolation in this path can be increased by the inclusion of ferrite unidirectional devices. Also care must be taken to provide equal length in both the test and reference paths. This often requires lengths of waveguide the same physical length as the device under test.

SWEPT FREQUENCY MIXER MEASUREMENTS

• Conversion Loss • Input VSWR • Ignores LO Feedthru



B/R = Conversion Loss/Flatness
A/R = Input VSWR

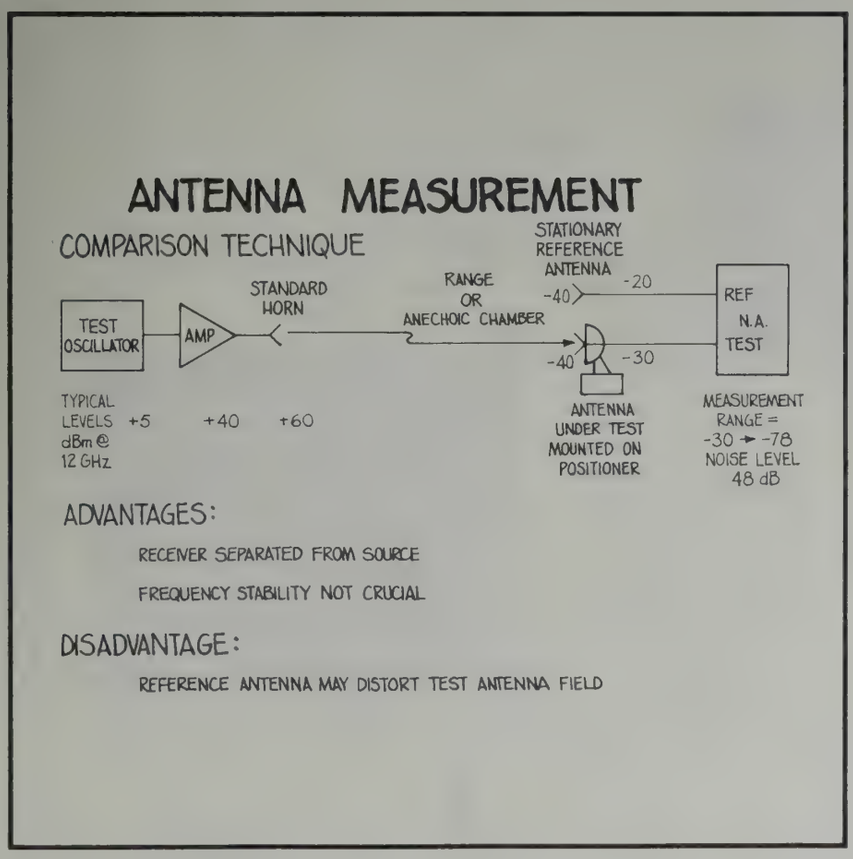
2 - 18 GHz RF
1- 300 MHz IF
Swept Measurement
1 Hz Offset Stability

APPLICATION NOTE 187-4

Designing and testing microwave mixers and receiver front-ends requires a local oscillator signal and the received signal, separated in frequency by the required IF. Broadband testing of such devices may be accomplished by taking CW measurements at discrete frequencies after manually tuning each of two signal sources. This method is both tedious and time-consuming, and it may result in the oversight of narrowband responses which occur between the discrete frequency points.

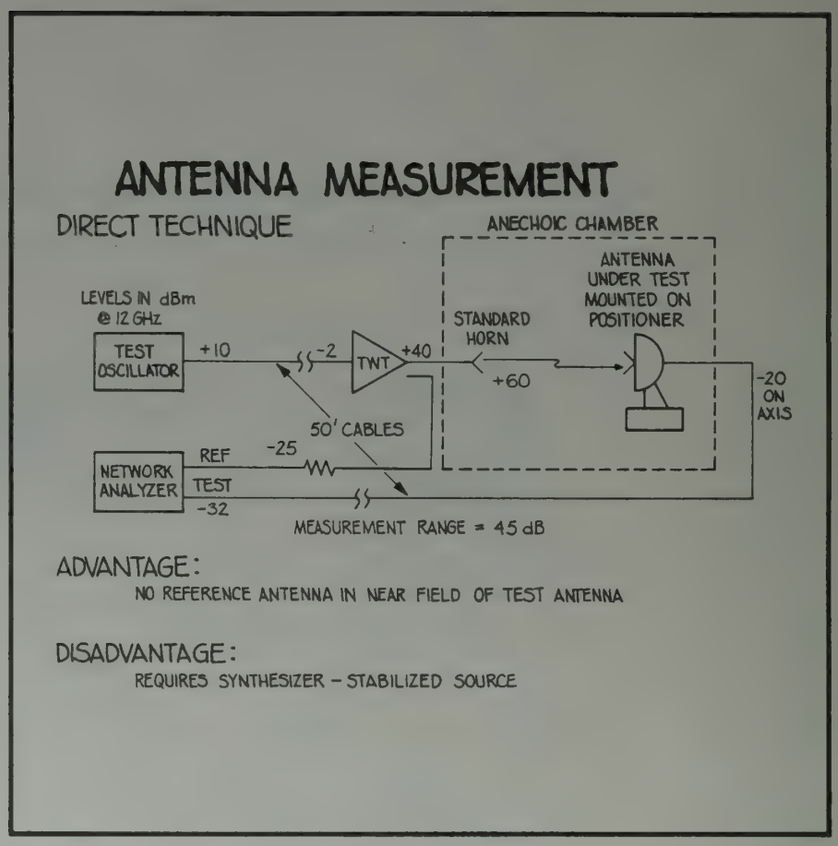
Application Note 187-4 describes the system shown above which will produce the two required signals, offset from each other by a very accurate and stable difference. The system can then be used to *sweep-test* across multi-octave bands. The operator needs only to set the controls of one sweeper (the "Master") to the desired frequency range between 2 and 18 GHz; the other sweeper will follow and always be separated by the desired offset, which can be anywhere between 1 and 300 MHz.

The Phase-Lock Controller provides RF and control circuitry to drive the "slave" sweeper and phase-lock it such that the offset between the two sweepers (the IF) is stable while they are sweeping.

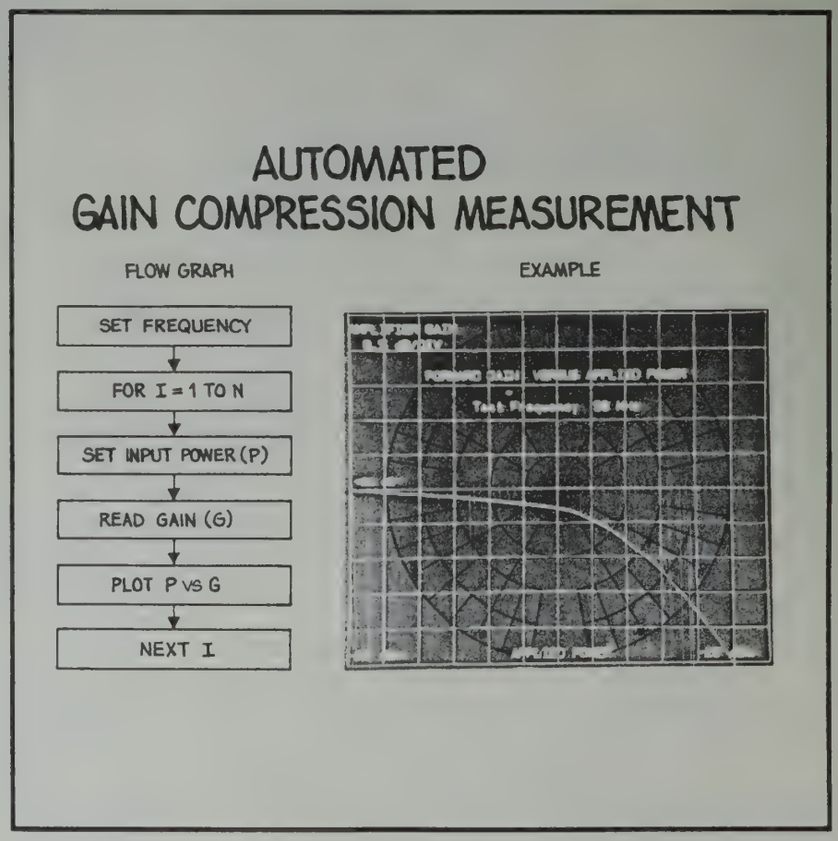


This is an example of antenna measurement using the comparison technique. If we are testing large aperture antennas we often have to employ a long outdoor antenna range to achieve far field measurement conditions. Amplifiers are necessary to overcome path and cable losses. We employ a separate antenna to establish a reference signal for our network analyzer. In spite of the great lengths involved the reference and test path lengths are comparable which minimizes the demands for source stability. This use of a separate antenna is suitable for higher gain antennae but with low gain antennae it might affect measurements.

Fortunately, low gain antennae don't require long ranges so we can then use an anechoic chamber and derive our reference signal from a coupler. However, since the test channel path is much longer than the reference channel, the phase response varies extremely rapidly vs. frequency. Therefore, a high stability source is required. The device under test is often not a single antenna but an array of antennae with and without a radome mounted in front. Since data is often required vs azimuth and elevation and polarization and frequency between each of 5 - 8 broadband spiral antennae, an automatic system is often essential for measurement and for logging the data in a format (mag tape or disc) suitable for input into a larger computer for further analysis and reduction.

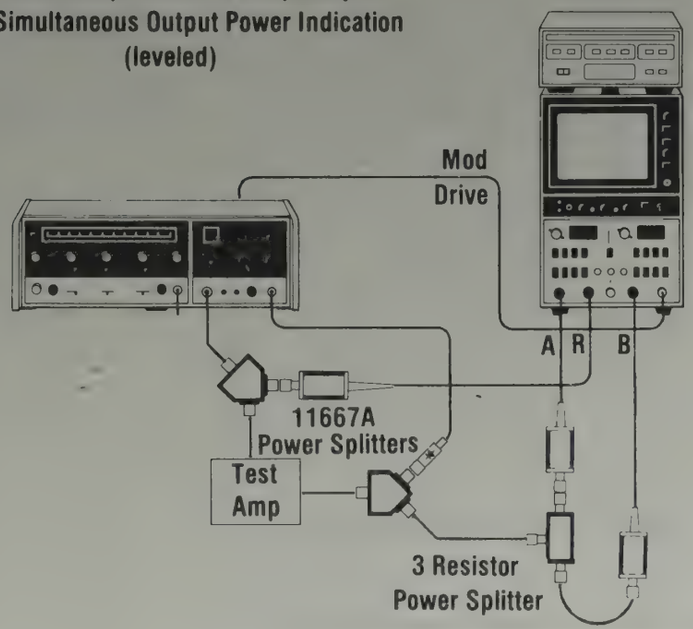


A common measurement required on amplifiers is gain compression or the variation in S_{21} as input power level is varied. Here is the output from an automatic system using the 8505 Network Analyzer in which the drive power is programmed in .1 dB steps and plotted versus the resultant gain. The labeled plot is generated by the graphics capability of the 8501 normalizer. This plot was performed with frequency held constant. However, a family of curves at many frequencies can be readily produced by a computer controlled system.

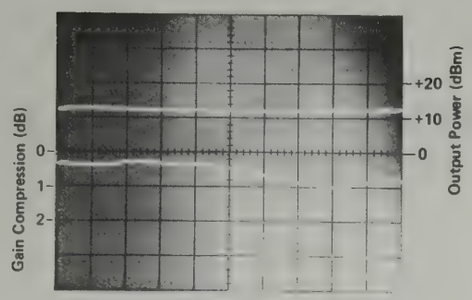


SWEPT FREQUENCY GAIN COMPRESSION

- Gain Compression vs Frequency
- Simultaneous Output Power Indication (leveled)



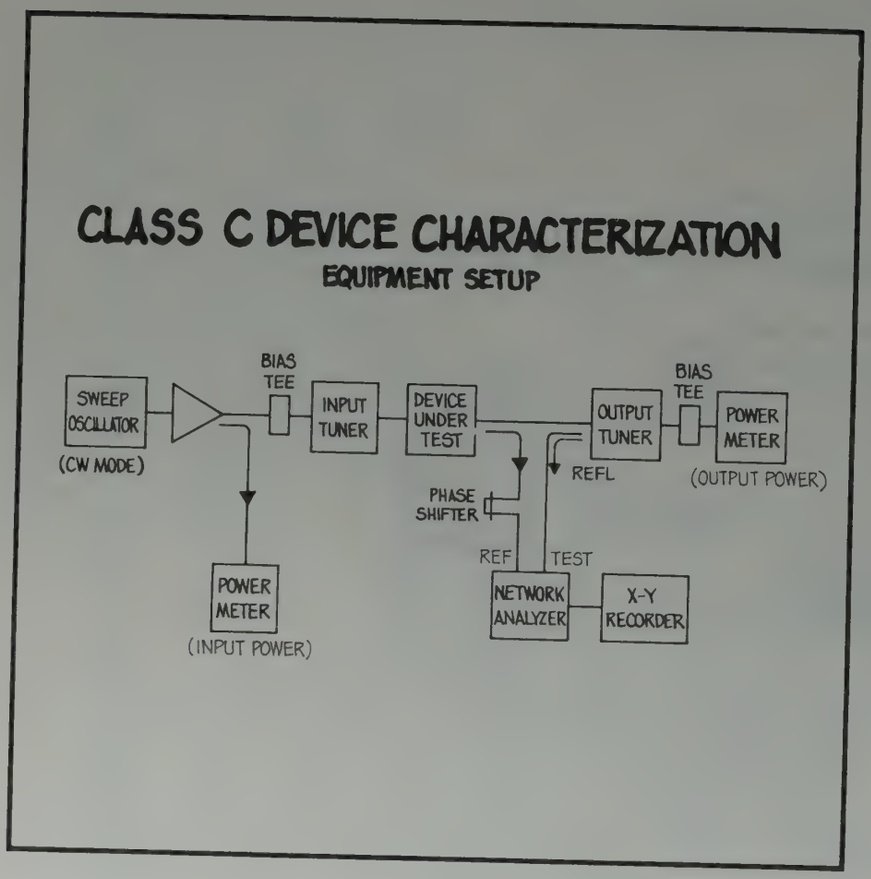
A/R = Gain Compression
B = Output Power



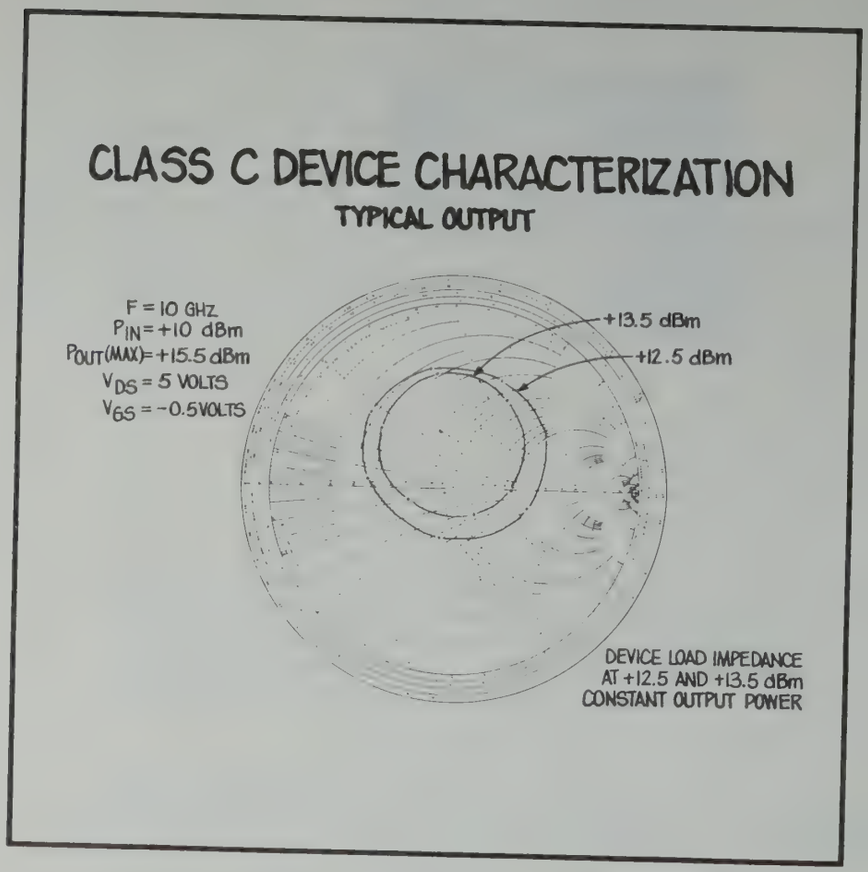
Application Note 155-1

If we could view gain compression on a swept basis, it would greatly speed tuning and testing. With this simple set-up, it is possible to observe gain-compression as well as power output versus frequency. A/R calibration is made using small-signal (low-level) gain. The 8750A is used to store and normalize this calibration. The power input is then increased until the gain drops 1 dB anywhere in the band of interest. The amplifier's output power can simultaneously be viewed on the B channel.

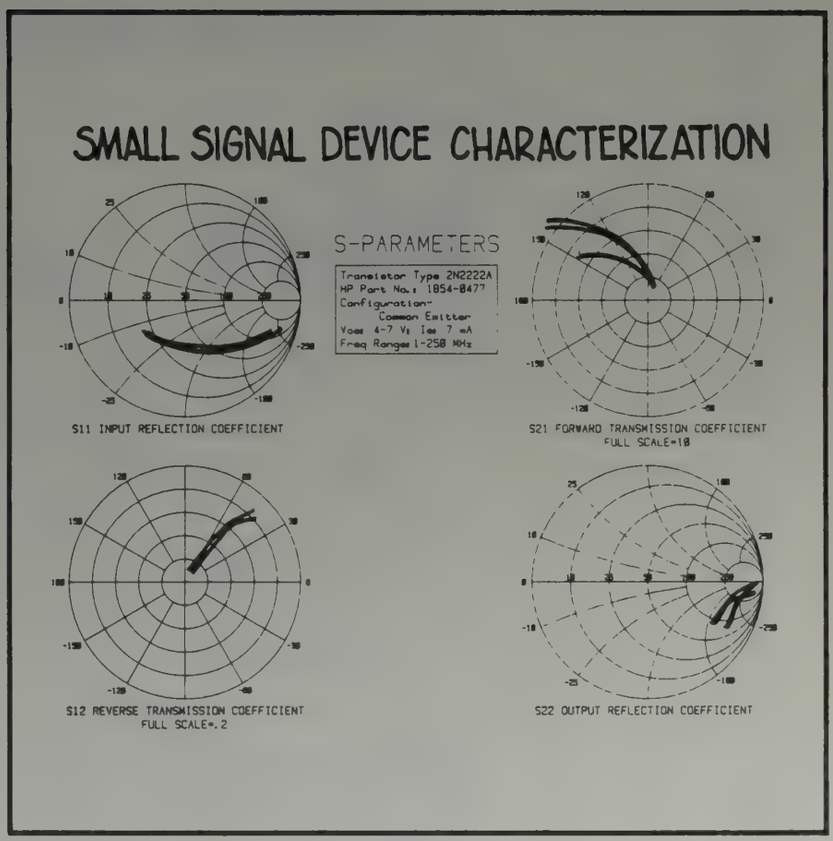
Since gain compression is an indication of the maximum output power possible before gain distortion occurs at any point in the band, it is necessary to level at the *output* of the amplifier. This compensates for the gain frequency response of the amplifier and ensures that the first point of compression is correctly shown.



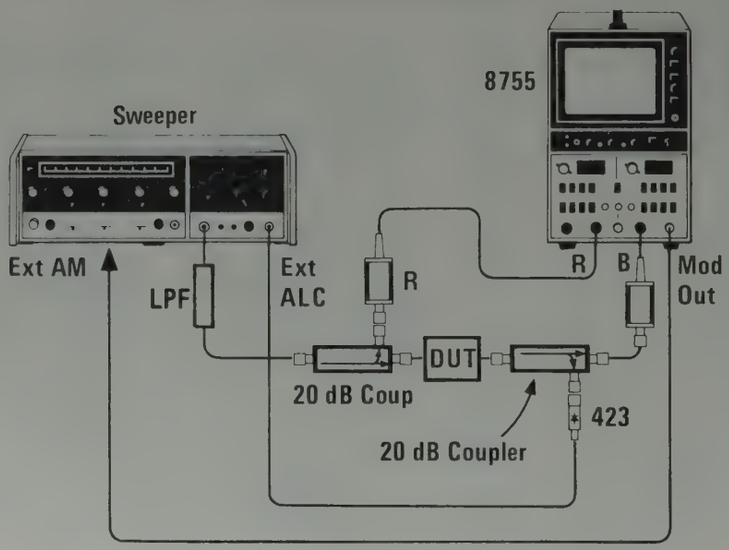
Another nonlinear application is the characterization of the output impedance required by a Class C transistor to develop a given output power. The circles are the locus of impedances required to provide a specific constant output power. The single point is the load impedance where maximum output power is developed for a given frequency and input power. The phase shifter is provided to allow calibration at the reference plane of the device under test. This block diagram refers to manually operated system in which one carefully adjusts the stub tuner to maintain a constant output power circle. However, this measurement is a likely candidate for automation if a computer controllable tuner is available.



One of the most common applications of the network analyzer is measurement of the Class A (small signal) S-parameters of a transistor. This output was generated by a controller-based system in which bias can be programmed and the switching between S-parameters is completely automatic.



INCREASED DYNAMIC RANGE USING HIGH POWER SOURCE



- With 86242C (5.9 – 9 GHz) or 86250C (8 – 12.4 GHz) → ~ 70 dB dynamic range

+17 dBm Source

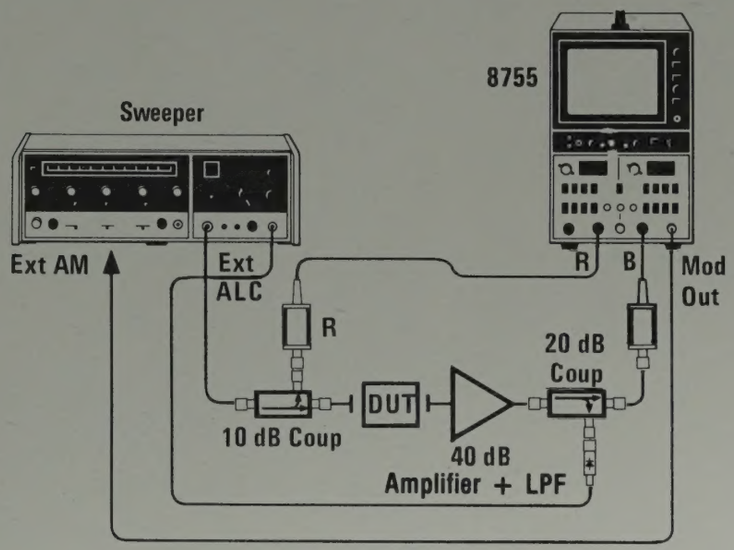
DUT ATTENUATION	R-DET LEVEL	B-DET LEVEL	SWEEPER P_0
0 dB (Calibr)	–10 dBm	+10 dBm	+10 dBm
7 dB	–3 dBm	↓	+17 dBm
10 dB	↓	+7 dBm	↓
20 dB		–3 dBm	
40 dB		–23 dBm	
60 dB		–43 dBm	
67 dB		–50 dBm	

We have been implying up to now that the dynamic range of the 8755 is limited to 60 dB (-50 dBm to + 10 dBm). Actually, this is the dynamic range of each detector. If the R detector is held constant in a ratio measurement, then indeed the dynamic range of the 8755 *is* limited to 60 dB. But if we were to somehow be able to utilize the dynamic range of the R detector, we could theoretically double dynamic range.

There are a couple of ways to partially accomplish this. One is shown above. There are two key requirements for this method: 1) the source needs to produce substantially more than 10 mW, and 2) we must externally level at the *output* of the DUT. During calibration the power level is set to provide 10 mW (+ 10 dBm) to the B detector and the leveling loop will attempt to hold this level constant; the level at the R detector during calibration is thus -10 dBm (see chart). When the attenuation of the DUT is 7 dB, the leveling detector calls for more power from the sweeper in its attempt to keep + 10 dBm at the B detector. A + 17 dBm source will be able to provide this power which will increase the level seen by R to -3 dBm (-10 + 7). Since the 8755 is displaying A/R, increasing the "R" level has the same effect as decreasing "B", i.e. in this example a 7 dB loss is displayed even though the power level at the B detector has not changed. For DUT's of greater attenuation, a + 17 dBm source will not be able to satisfy the leveling loop and the level at the B detector will begin to decrease down to the noise level of -50 dBm. This means a dynamic range of > 67 dB is possible using a + 17 dBm source such as the HP 86242C (5.9-9 GHz) or 86250C (8-12.4 GHz) plug-ins. Typically the dynamic range of such a combination would be more than 70 dB.

To obtain much more dynamic range with this method would require very high power sources, and appropriate precautions would have to be taken (e.g. 90 dB dynamic range would require a 10 Watt source).

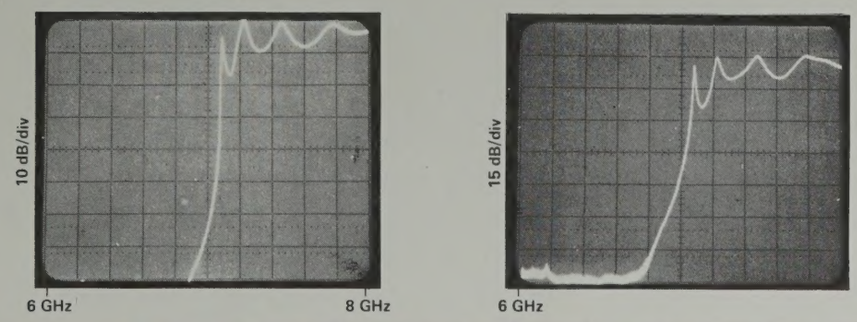
INCREASED DYNAMIC RANGE USING AN EXTERNAL AMPLIFIER



● > 100 dB dynamic range (with proper amplifier)

+10 dBm Source

DUT ATTENUATION	R-DET LEVEL	B-DET LEVEL	SWEEPER P_o
0 dB (Calibr)	–40 dBm	+10 dBm	–30 dBm
20 dB	–20 dBm	↓	–10 dBm
40 dB	0 dBm	↓	+10 dBm
60 dB	↓	–10 dBm	↓
80 dB		–30 dBm	
100 dB		–50 dBm	



To obtain as much as 100 dB dynamic range with the 8755, this set-up could be used. Here a similar concept is employed (i.e., use some of the R detector's dynamic range), but by placing an amplifier *after* the device under test, its critical characteristic is high gain not high power. Again the calibration power level is set such that the B detector sees + 10 dBm and the leveling loop attempts hold this level constant. The table shows the effects of various DUT attenuators on the detectors and source. The important point is that only a + 10 dBm source is needed to achieve a 100 dB dynamic range when a 40 dB gain amplifier is available.

Because of noise considerations and leveling loop range, 100 dB is pretty much the limit of this set-up even with a higher gain amplifier. It would be even less (~20 dB less) if we used an unmodulated detection scheme because of the broadband noise contributed by the high-gain amplifier. Of course, the plug-in used must be capable of being directly modulated by the 8755; otherwise the leveling loop would attempt to buck the modulation signal.